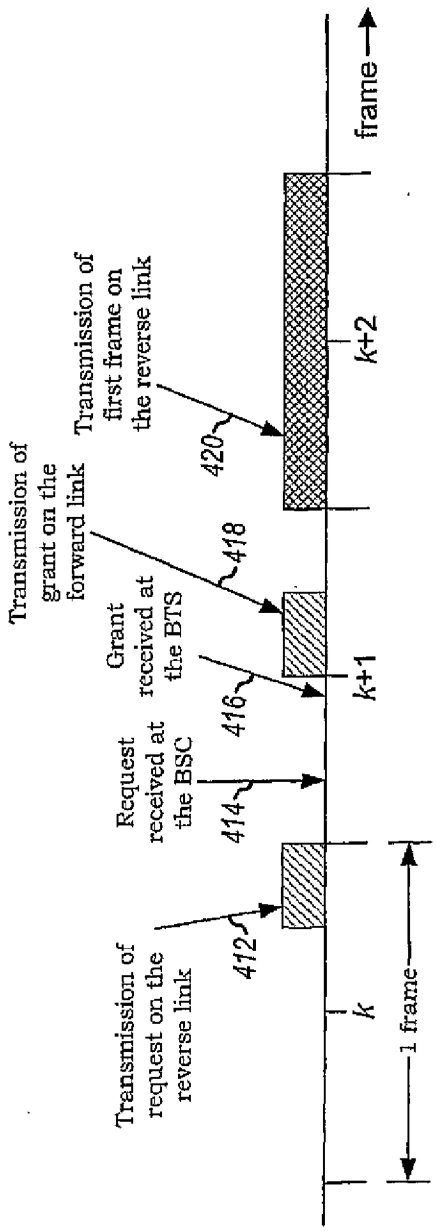
Transmission of request on the reverse link
412
Request received at the BSC
414
Grant received at the BTS
416
Transmission of grant on the forward link
418
Transmission of first frame on the reverse link
420
k
k+1
k+2
1 frame
frame

FIG. 4

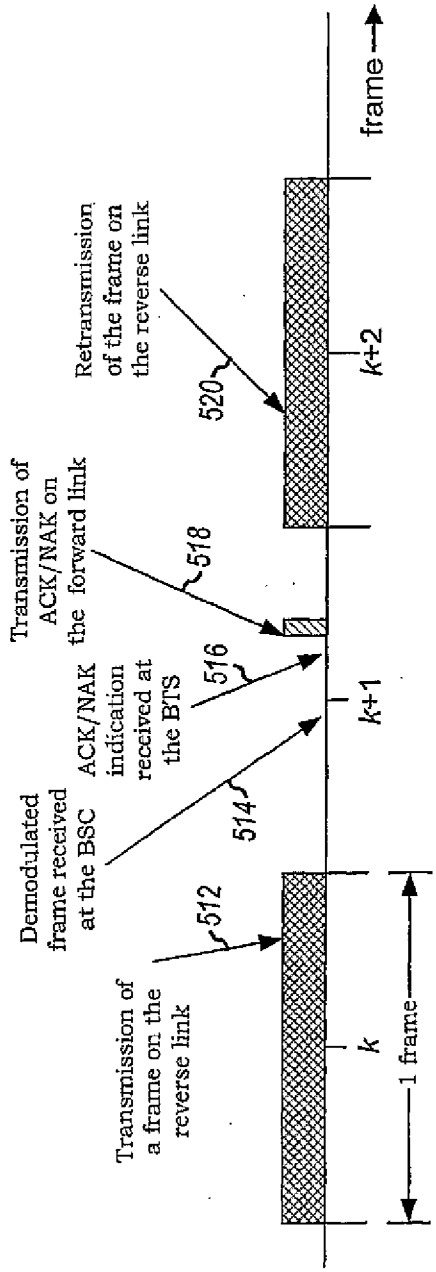
Transmission of a frame on the reverse link
512
Demodulated frame received at the BSC
514
ACK/NAK indication received at the BTS
516
Transmission of ACK/NAK on the forward link
518
Retransmission of the frame on the reverse link
520
k
k+1
k+2
1 frame
frame

FIG. 5A

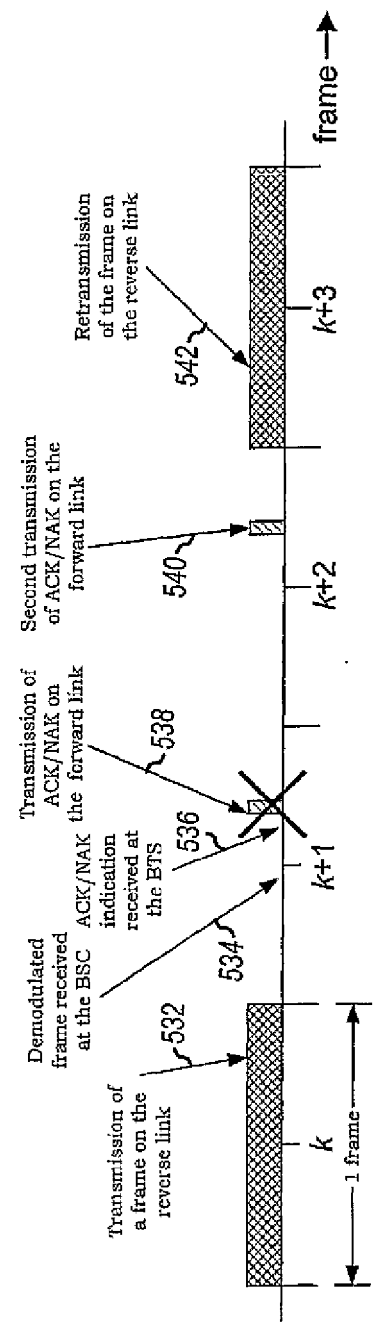
Transmission of a frame on the reverse link
532
Demodulated frame received at the BSC
534
ACK/NAK indication received at the BTS
536
Transmission of ACK/NAK on the forward link
538
Second transmission of ACK/NAK on the forward link
540
Retransmission of the frame on the reverse link
542
k
k+1
k+2
k+3
1 frame
frame

FIG. 5B

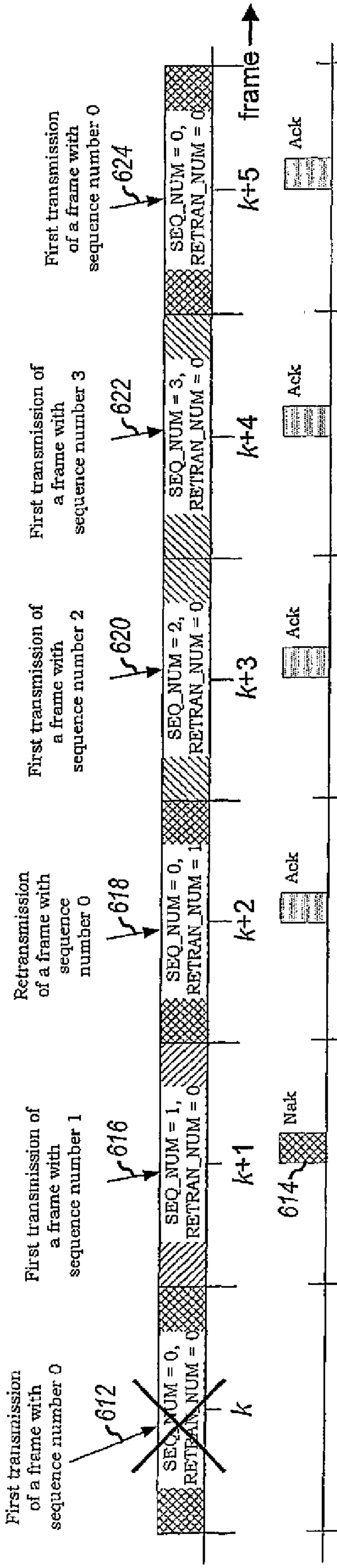
First transmission of a frame with sequence number 0
612
SEQ_NUM = 0, RETRAN_NUM = 0
k
First transmission of a frame with sequence number 1
616
SEQ_NUM = 1, RETRAN_NUM = 0
k+1
614
Nak
Retransmission of a frame with sequence number 0
618
SEQ_NUM = 0, RETRAN_NUM = 1
k+2
Ack
First transmission of a frame with sequence number 2
620
SEQ_NUM = 2, RETRAN_NUM = 0
k+3
Ack
First transmission of a frame with sequence number 3
622
SEQ_NUM = 3, RETRAN_NUM = 0
k+4
Ack
First transmission of a frame with sequence number 0
624
SEQ_NUM = 0, RETRAN_NUM = 0
k+5
Ack
frame

FIG. 6A

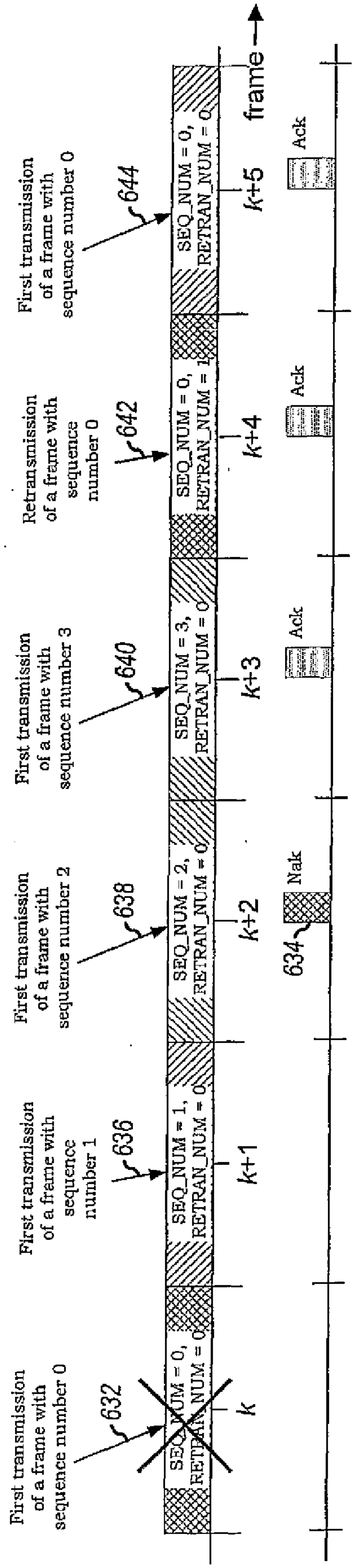
First transmission of a frame with sequence number 0
632
SEQ_NUM = 0, RETRAN_NUM = 0
k
First transmission of a frame with sequence number 1
636
SEQ_NUM = 1, RETRAN_NUM = 0
k+1
First transmission of a frame with sequence number 2
638
SEQ_NUM = 2, RETRAN_NUM = 0
k+2
634
Nak
First transmission of a frame with sequence number 3
640
SEQ_NUM = 3, RETRAN_NUM = 0
k+3
Ack
Retransmission of a frame with sequence number 0
642
SEQ_NUM = 0, RETRAN_NUM = 1
k+4
Ack
First transmission of a frame with sequence number 0
644
SEQ_NUM = 0, RETRAN_NUM = 0
k+5
Ack
frame

FIG. 6B

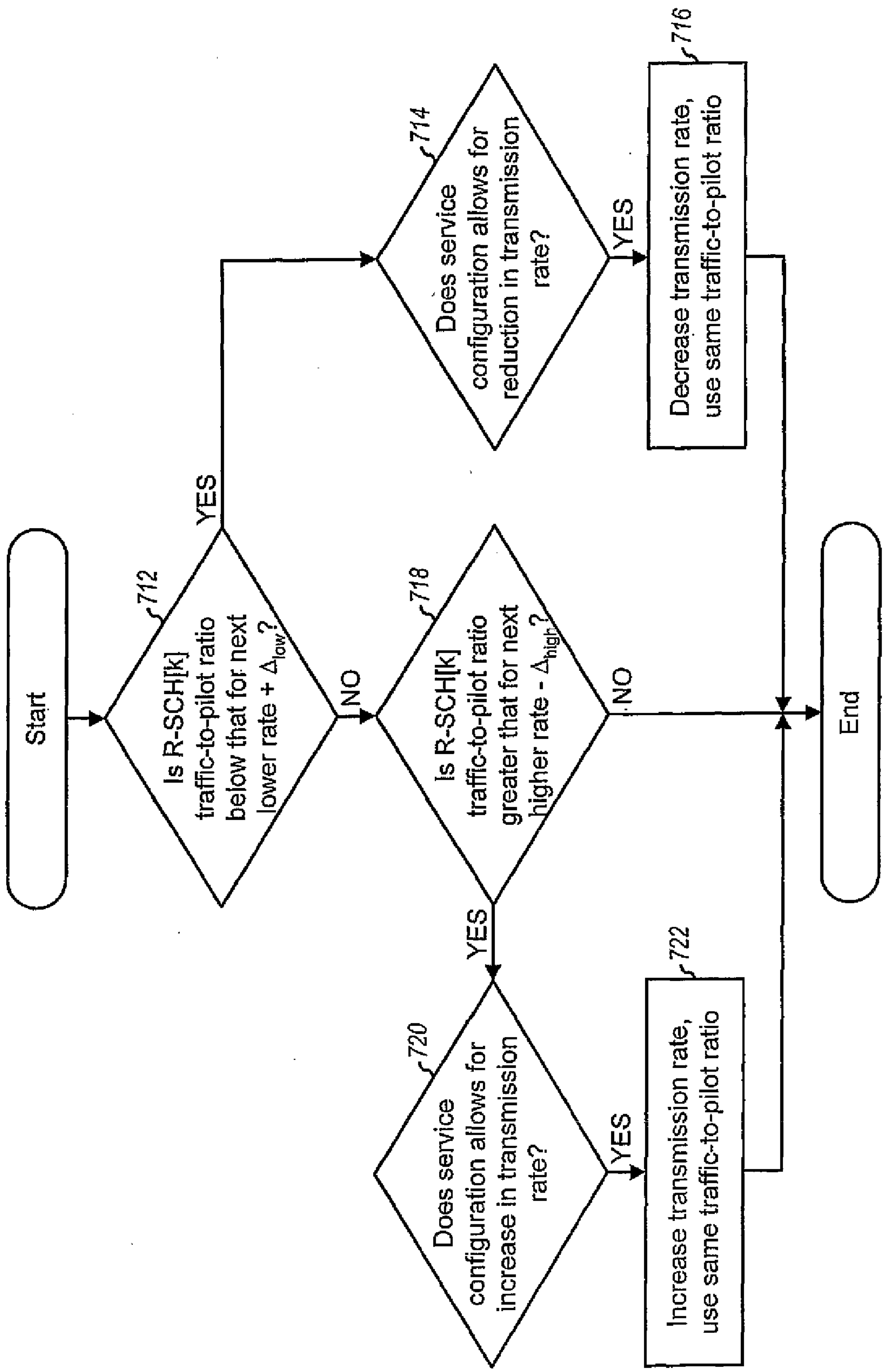
Start
712
Is R-SCH[k] traffic-to-pilot ratio below that for next lower rate + Δ_{low}?
YES
NO
714
Does service configuration allows for reduction in transmission rate?
YES
716
Decrease transmission rate, use same traffic-to-pilot ratio
718
Is R-SCH[k] traffic-to-pilot ratio greater that for next higher rate - Δ_{high}?
NO
YES
720
Does service configuration allows for increase in transmission rate?
YES
722
Increase transmission rate, use same traffic-to-pilot ratio
End

FIG. 7

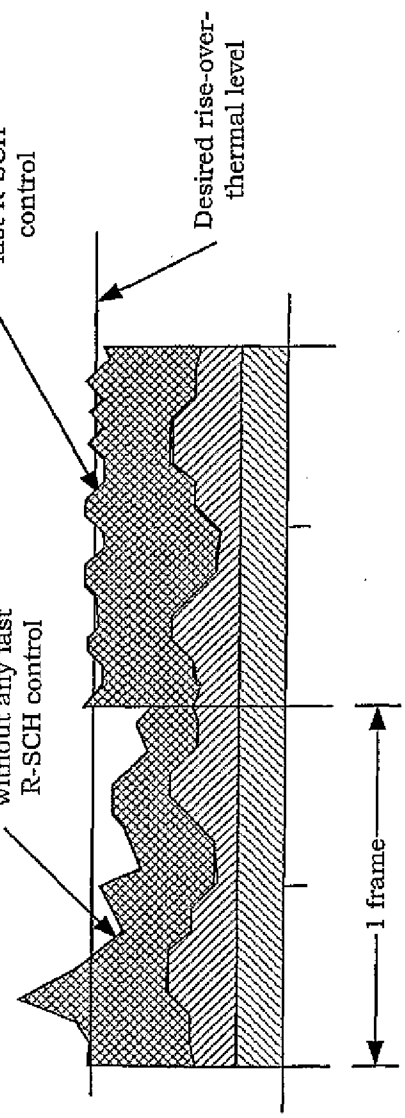
Power level with
fast R-SCH
control
Desired rise-over-
thermal level
Power level
without any fast
R-SCH control
1 frame

FIG. 8

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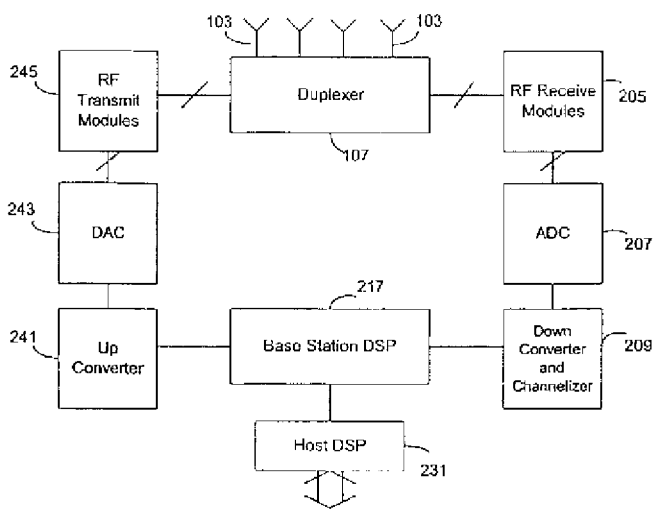
(54) Title: FREQUENCY DEPENDENT CALIBRATION OF A WIDEBAND RADIO SYSTEM USING NARROWBAND CHANNELS



(57) Abstract: A method and apparatus are provided that determine group delay for a set of transmit or receive chains over a wide frequency band without causing significant interference with simultaneous users of the system. In one embodiment, the invention includes an antenna array adapted to transmit and receive radio communications signals with a plurality of other terminals, the communications signals each using a particular minimum bandwidth, a transmit chain to transmit a calibration signal through the antenna array to a transponder on at least two different frequency bands within the minimum bandwidth, and a receive chain to receive through the antenna array a transponder signal from the transponder, the transponder signal being received on at least two different frequency bands and being based on the calibration signal. A signal processor determines a frequency dependent calibration vector based on the at least two frequency bands of the transponder signal as received through the receive chain.

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For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

FREQUENCY DEPENDENT CALIBRATION OF A WIDEBAND RADIO SYSTEM USING NARROWBAND CHANNELS

BACKGROUND OF THE INVENTION

Field of the Invention

The invention relates generally to the field of digital signal communications and to receive and transmit chain calibration. More particularly, the invention relates to calibrating the group delay using narrowband signals at more than one frequency.

Description of the Related Art

Radio communications capacity can be greatly increased using directional, rather than omni-directional radio transmission. One way to transmit directional signals and directionally receive signals is by using beam forming and nulling through an array of antennas. The precision of the beam forming and nulling through the antenna array, can be improved if the transmit and receive chains are both calibrated. Calibration can be applied to the chain from the digital interface at baseband to the field radiated from or received at each antenna element. One way of making the calibration is to have a transponder separated from the antenna array listen to the output of the antenna array on a base station downlink frequency. The transponder receives a downlink calibration signal from the base station and then re-transmits it on an uplink frequency. By selecting appropriate signals to transmit and appropriate signals to receive, the base station can apply signal processing to estimate compensations in phase and amplitude to calibrate its transmit and receive chains.

A remote transponder calibration system is shown, for example, in U.S. Patent No. 5,546,090 to Roy, III et al. That patent describes calibrating a narrowband FDD (frequency division duplex) system for phase and amplitude at each transmit and receive chain. In an FDD system, unused time and frequency slots typically occur on occasion and these can be used to send and receive a narrowband calibration signal. In a typical spread spectrum system, however, there are no unused time and frequency slots to use for calibration. A spread spectrum system, for example a CDMA (code division multiple access) system, as opposed to FDMA (frequency division multiple access) and TDMA (time division multiple access) systems, has multiple users using the same radio channel at the same time. If the transponder is designed to receive and transmit the signal using the same spread spectrum channel that is used for traffic, then the additional energy added to the channel by calibration will reduce system

capacity. A typical transponder will receive all of the downlink traffic including the calibration signal, shift the frequency, amplify it and send all of the traffic back to the base station. This results in a very large amount of energy being sent by the transponder on the uplink and may effectively overpower all other traffic. As a result, calibration will affect both the downlink and uplink channel capacity. For calibrating group delay for a set of transmitters or receivers, a calibration signal normally is transmitted across a wider band of frequencies further ensuring interruptions to normal traffic.

BRIEF SUMMARY OF THE INVENTION

A method and apparatus are provided that determine group delay for a set of transmit or receive chains over a wide frequency band without causing significant interference with simultaneous users of the system. In one embodiment, the invention includes an antenna array adapted to transmit and receive radio communications signals with a plurality of other terminals, the communications signals each using a particular minimum bandwidth, a transmit chain to transmit a calibration signal through the antenna array to a transponder on at least two different frequency bands within the minimum bandwidth, and a receive chain to receive through the antenna array a transponder signal from the transponder, the transponder signal being received on at least two different frequency bands and being based on the calibration signal. A signal processor determines a frequency dependent calibration vector based on the at least two frequency bands of the transponder signal as received through the receive chain.

Other features of the present invention will be apparent from the accompanying drawings and from the detailed description that follows.

BRIEF DESCRIPTION OF THE SEVERAL VIEWS OF THE DRAWINGS

The present invention is illustrated by way of example, and not by way of limitation, in the figures of the accompanying drawings in which like reference numerals refer to similar elements and in which:

Figure 1 is a block diagram illustrating an exemplary architecture of a wireless communication system base station appropriate for use with one embodiment of the present invention;

Figure 2 is a block diagram illustrating an exemplary architecture of a wireless transponder system appropriate for use with the base station of Figure 1;

Figure 3 is a process flow diagram showing the calibration of a receive chain; and

Figure 4 is a process flow diagram showing the calibration of a transmit chain.

DETAILED DESCRIPTION OF THE INVENTION

Introduction

In one embodiment, the present invention includes a method for calibrating the group delay of multiple transmit and receive chains of a wideband adaptive antenna base station using a narrowband transponder. In order to calibrate the group delay of the transmit and the receive chains, the base station transmits a different narrowband calibration signal over each of the transmit chains on at least two different frequencies in the downlink frequency band. These signals are then received by the narrowband transponder and retransmitted to the base station as narrowband signals in the wideband uplink frequency band. In this application, the radios in the adaptive antenna base station support wideband channels. However, in order to avoid creating any unnecessary interference, the calibration signals and the transponder signals are narrowband. In other words, the calibration signals occupy only narrow portions of the wideband channel. The transponder only receives in these narrow frequency bands and only retransmits the signals in correspondingly narrow portions of the uplink band.

Since the narrowband signals add only a small amount of energy to the wideband uplink and downlink channels, the calibration can be done while regular data traffic is being supported by the base station. The narrower the bandwidth of the calibration signals, the less will be the amount of energy that will be added to the system. For wideband spread spectrum systems the narrowband signals can easily be one tenth, or one hundredth as wide as the regular data traffic channels. For frequency division systems, the narrowband signals can still be one third to one fifth the width of the traffic channels. Proper selection of the signal power levels can further reduce the impact on regular traffic. Using multiple narrowband signals and transponder bands it is possible to calibrate for more general phase and gain variations as a function of

frequency. In a CDMA (Code Division Multiple Access) system, it is possible to design the CDMA system to be particularly insensitive to narrowband signals.

In one embodiment, the transponder only receives and re-transmits on narrow bands within the traffic bands of the wider band system to be calibrated. The system can have a set of wideband transmitters with antenna elements and a set of wideband receivers with antenna elements or a single set of elements can be common to the transmitters and the receivers. In both cases, system performance is normally improved with frequent calibration of the group delay for both the transmit chain and the receive chain. The group delay calibration vectors can be different for the receive chain and the transmit chain. In one example, the system has a multi-channel base station that communicates with multiple subscribers up to 10km away using CDMA with SDMA (spatial division multiple access). For this system, it has been found that calibrations every hour or two will noticeably improve performance. With such frequent calibrations, the impact of calibration on normal operations can be important. According to the present invention, the impact of calibration on normal operations can be minimized with a narrowband calibration transponder.

On each narrow frequency calibration band, different signals can be transmitted through two or more transmit chains. The signals can be differentiated, for example, by modulating different sequences onto the signals. In one embodiment, the sequences are orthogonal sequences to aid in demodulation. In another embodiment, the sequences are modulated onto the signals as spreading codes. This allows de-spreading codes to be used on the received signal so that the signal from each transmit chain can be distinguished. The transponder receives these signals and re-transmits them in the base station uplink band. The signals received by the base station can then be processed in order to measure any desired relative characteristics of the signals. For example, the signals can be used to find the relative phase and amplitude of the involved transmit chains and the relative phase and amplitude of all the receive chains. By transmitting different signals over the different transmit chains, the signals can be differentiated when received. This allows characteristics such as relative phase and amplitude to be estimated separately for each transmit chain. The characteristics can be used to determine spatial signatures for the uplink and downlink as well as to calculate frequency dependent calibration vectors. Combining phase measurements at different frequencies, a group delay calibration vector can be derived.

The relative phase and amplitude of the transmit chains can be estimated by receiving the different signals at a single antenna and then estimating the channel for each of the different signals transmitted over the different transmit chains. The relative phase and amplitude of the receive chains can be estimated by transmitting a single calibration signal over a single transmit chain and receiving it over the different receive chains. The channel received over each receive chain can then be estimated and compared to find spatial signatures and for calibration. As a result, if the calibration signal is sent once over all transmit chains and then the corresponding transponder signal is received through all receive chains, the entire array can be calibrated based on a single downlink and uplink burst. Since the transmit and receive calibration vector determinations need not be coupled to each other, performing both calibrations on the same burst increases efficiency and reduces the effects on traffic. If the calibration signal is transmitted on two or more different frequencies either at the same time or at different times close together, then the group delay can be derived.

As an alternative, just a few or even two of the transmit or receive chains can be calibrated at one time. If all the transmit or receive chains are not involved in each calibration measurement, then repeated calibration measurements with different sets of transmit or receive chains can be performed so that all relative phases and amplitudes can be measured among all the transmit and receive antennas. Accuracy is improved if there is a common transmit or receive chain in each of the measurements. This allows the measured phases and amplitudes to be related to each other with reference to the common chain. Typically, one of the receive chains is designated as a reference receive chain and calibration signals are measured in pairs with each receive chain being paired with the reference chain. Since the reference chain participates in every measurement, all of the other chains can be referenced to each other through the reference chain. After the receive chains are calibrated, a similar process is performed with the transmit chains being measured in pairs against the reference. It is not important which particular chain is selected to be the reference and the receive and transmit references need not have any relationship to each other. The calibration vectors can be expressed as variations from the reference or from any arbitrary standard such as an average, mean, or median of the differences between the receive or transmit chains, respectively.

In one embodiment, the present invention is implemented in an SDMA radio data communications system. In such a spatial division system, each terminal is

associated with a set of spatial parameters that relate to the radio communications channel between, for example, the base station and a user terminal. The spatial parameters comprise a spatial signature for each terminal. Using the spatial signature and arrayed antennas, the RF energy from the base station can be more precisely directed at a single user terminal, reducing interference with and lowering the noise threshold for other user terminals. Conversely, data received from several different user terminals at the same time can be resolved at lower receive energy levels. With spatial division antennas at the user terminals, the RF energy required for communications can be even less. The benefits are even greater for subscribers that are spatially separated from one another. The spatial signatures can include such things as the spatial location of the transmitters, the directions-of-arrival (DOAs), times-of-arrival (TOAs) and the distance from the base station.

Estimates of parameters such as signal power levels, DOAs, and TOAs can be determined using known training sequences placed in digital data streams for the purpose of channel equalization in conjunction with sensor (antenna) array information. This information is then used to calculate appropriate weights for spatial demultiplexers, multiplexers, and combiners. Extended Kalman filters or other types of linear filters, well known in the art, can be used to exploit the properties of the training sequences in determining spatial parameters. Further details regarding the use of spatial division and SDMA systems are described, for example, in U.S. Patents Nos. 5,828,658, issued Oct. 27, 1998 to Ottersten et al. and 5,642,353, issued June 24, 1997 to Roy, III et al.

Base Station Structure

The present invention relates to wireless communication systems and may be a fixed-access or mobile-access wireless network. It may use spatial division technology in combination with wideband multiple access systems, such as code division multiple access (CDMA), and other spread spectrum type systems. Figure 1 shows an example of a base station of a wireless communications system or network suitable for implementing the present invention. The system or network includes a number of subscriber stations, also referred to as remote terminals or user terminals, (not shown). The base station may be connected to a wide area network (WAN) through its host DSP 231 for providing any required data services and connections external to the immediate wireless system. To support spatial division, a plurality of

antennas 103 is used, for example four antennas, although other numbers of antennas may be selected.

The outputs of the antennas are connected to a duplexer switch 107, which in this CDMA system is a frequency switch. Alternatively, separate transmit and receive antenna arrays can be used, in which case the duplexer is not necessary. When receiving, the antenna outputs are connected via the switch 107 to RF (radio frequency) receive modules 205, and are mixed down and channelized in a down converter 207. The down converted signals are then sampled and converted to digital in an ADC (analog to digital converter) 209. This can be done using FIR (finite impulse response) filtering techniques. The invention can be adapted to suit a wide variety of RF and IF (intermediate frequency) carrier frequencies and bands.

There are, in the present example, four antenna channel outputs, one from each antenna receive module 205. The particular number of channels can be varied to suit network needs. For each of the four receive antenna channels, the four down-converted outputs from the four antennas are fed to a digital signal processor (DSP) device 217 for further processing, including calibration. According to one aspect of this invention, four Motorola DSP56300 Family DSPs can be used as channel processors, one per receive channel. The timeslot processors 217 monitor the received signal power and estimate the phase and time alignment. They also determine smart antenna weights for each antenna element. These are used in the spatial division multiple access scheme to determine a signal from a particular remote user and to demodulate the determined signal.

The output of the channel processors 217 is demodulated burst data. This data is sent to the host DSP 231 whose main function is to control all elements of the system and interface with the higher level processing. The higher level processing provides the signals required for communications in all the different control and service communication channels defined in the system's communication protocols. The host DSP 231 can be a Motorola DSP56300 Family DSP. In addition, channel processors send the determined receive weights for each user terminal to the host DSP 231.

The host DSP 231 maintains state and timing information, receives uplink burst data from the channel processors 217, and programs the channel processors 217. In addition, it decrypts, descrambles, checks error detecting code, and deconstructs bursts of the uplink signals, then formats the uplink signals to be sent for higher level processing in other parts of the base station. With respect to the other parts of the

base station, it formats service data and traffic data for further higher processing in the base station, receives downlink messages and traffic data from the other parts of the base station, processes the downlink bursts and formats and sends the downlink bursts to the transmit chain, discussed below.

Transmit data from the host DSP 231 is used to produce analog transmit outputs which are sent to the RF transmitter (tx) modules 245. Specifically, the received data bits are converted via a DAC (digital to analog converter) 241 to analog transmit waveforms and up-converted into a complex modulated signal, at an IF frequency in an upconverter 243. The analog waveforms are sent to the transmit modules 245. The transmit modules 245 up-convert the signals to the transmission frequency and amplify the signals. The amplified transmission signal outputs are sent to antennas 103 via the duplexer/time switch 107.

Narrowband Transponder Structure

Referring to Figure 2, an example of a remote transponder, suitable for use in implementing the present invention is shown. This transponder is designed to be inexpensive and simple. The particular transponder design shown can also be made in a small, portable, and lightweight package that can be used at the installation of the base station, if desired. The transponder can be mounted on a nearby fixture or even on the antenna mast that is used by the base station's antennas. Alternatively, the transponder can instead be operated as a special mode of a much more complex and fully functional user terminal. A second base station can also perform the transponder functions. The function of the transponder 118 is to receive a signal in the range of the wideband downlink channel, up-convert or down-convert it to the wideband uplink channel, filter it to select only a narrow frequency band, amplify it, and then re-transmit it as a signal in the range of the uplink channel. As mentioned above, frequency-shifting transponder 118 is only one possible example of a transponder suitable for use in calibration. The only general requirement for the transponder is that it transmits back a radio frequency signal that is somehow distinguishable from the signal it received. Besides frequency shifting the signal, the transponder can also time delay the signal, or more generally modulate it with various well-known modulation schemes. For a code division multiplex system, the transponder can also decode the received signal and encode it with a new spreading code for the uplink channel.

As shown in Figure 2, the calibration signal from the base station is received at the transponder antenna 122. A duplexer 140 separately routes signals received at the antenna to the receive chain beginning with a receive bandpass filter 126 and signals coming from the transmit chain, ending with a transmit bandpass filter 125. In the receive chain, signals coming from the transponder antenna after filtering 125 are routed to a low noise amplifier (LNA) 142. This amplified signal is then filtered again by a bandpass filter 144, which eliminates unwanted signals based on their frequencies. This filtered signal is then down-converted to IF (intermediate frequency) by a mixer 148 that combines the received signal with a LO (local oscillator signal) 146 waveform. The IF signal is processed through another bandpass filter 150 before upconversion for transmission. The channel filter 150 can be configured to have two or more passbands, one for each of the frequencies of the calibration signal from the base station.

A second mixer 149 combines the signals from the bandpass filter 150 and a second LO 147 to produce two new transmit signals at frequencies spaced apart from each other and within the uplink frequency band. These two new signals are bandpass filtered 145 and amplified in a power amplifier 143. The power amplifier is adjusted by a power feedback control loop 141 to reduce interference with other channels and smooth reception of the calibration signal at the base station. Another bandpass filter 125 eliminates the upper mixer product and any artifacts from the power amplifier, leaving only the lower mixer product which is a copy of the original input signal on the RF receive chain except for its frequency. This signal is connected to the duplexer 140 for transmission through the antenna element 122. The transponder shows, as an alternative, a separate transmit antenna element 123 and receive antenna element 124. If separate elements are used then the duplexer 140 is no longer required and the antennas can be directly coupled to the respective transmit and receive bandpass filters.

The transponder described above is designed to shift and transpond narrowband signals from the base station that are transmitted in the band for North American cellular CDMA communications, designated as IS-95 by the Telecommunications Industry Association (TIA). In some circumstances, it might be desirable to receive a wideband calibration signal over the complete CDMA channel and return it as a narrowband signal. Since most single channel communication bandwidths are too wide for practical filters at RF frequencies, such a single channel transponder would mix the RF frequency down to a lower intermediate frequency,

apply a narrowband filter at this intermediate frequency, and then mix the filtered signal back up to the desired RF frequency to be echoed back as a narrowband signal. In all other aspects, the wideband, single channel, transponder would behave and be constructed like the narrowband transponder described here.

To determine group delay, at least two frequencies of the calibration signal are desired. To return the two frequencies of the calibration signal, the transponder can be configured to return the two narrowband signals shifted in frequency. Alternatively an additional transponder with unique or some shared hardware can be used. Each transponder can be configured to receive and transmit only in a narrow band or to receive and transmit a broad range of different frequencies. The particular design of the multiple frequency transponder system will depend on the particular circumstances of the application and the communication system.

In operation, the base station DSP 217 generates a specialized narrowband calibration transmit signal on at least two frequencies which it transmits from the antenna array through the duplexer. The transponder receives the calibration transmit signal and echoes it back with the appropriate changes so that it will be received through the receive chain through the duplexer. In a conventional cellular CDMA system, the radio system uses different frequencies for transmit and receive. Thus, the transponder echoes back a signal on the uplink frequency band that is a frequency-shifted copy of the downlink signal it receives. The base station DSP acquires the echoed calibration signal on both frequencies through the receive chain and uses this received calibration signal along with knowledge of the transmit calibration signal to calculate group delay vectors which are then stored in a group delay calibration vector storage buffer.

For a CDMA cellular system, the system may be allocated a bandwidth from, e.g., 824 MHz to 835 MHz or from 835MHz to 849 MHz. The wideband channels within this range may be as narrow as 1.25 MHz or as wide as 5 MHz. In such a system, uplink and downlink frequency bands are typically separated from each other with a significant guard band so that they are separated by 1.25 MHz to 5 MHz. This is the amount by which the transponder must shift the calibration signal frequency to send it back to the base station. In other systems, the wideband uplink and downlink channels may be as wide as 40 MHz or more. The narrowband calibration signals on the other hand, would typically be from 0.01 MHz to 0.1 MHz wide. The spectral width of the calibration signal will be as small as reasonably convenient with readily available equipment at moderate cost. The narrower the signal, the less it will

interfere with existing traffic. However, as mentioned above, the narrowband signal must also be able to be transmitted and received by the wideband transmit and receive chains. The necessary bandwidth limitations will also depend on the particular system. For a system in which the wideband signals are 1.25 MHz wide, the narrowband signals will probably be much narrower than for a system in which the wideband signals are 40 MHz wide. The particular carrier frequencies used can also be adapted to suit the needs of the particular system. Currently, appropriate systems have carrier frequencies centered at frequencies ranging from 450 MHz to 2100 MHz. This range is expected to become greater as radio technologies and spectrum allocations change.

Calculation of Calibration Vectors

There are a variety of different ways to calculate and calibrate the phases and amplitudes of a multiple antenna array using narrowband signals and a transponder. U.S. Patents Nos. 5,546,090 issued August 13, 1996 to Roy, III et al., 5,930,243 issued July 27, 1999 to Parish et al. and 6,037,898 issued to Parish et al. show suitable approaches to calibration. Another approach is shown in International Application No. WO99157820, published November 11, 1999 of Boros et al. The disclosures of these references are hereby incorporated by reference herein.

With respect to calibrating the group delay for the transmit and receive chains of the base station, assuming identical RF propagation on the uplink and downlink, a single transponder or subscriber unit can be used together with its base station to carry out the calibration. However, the present invention enables the separate determination of the uplink and downlink signatures for the transponder or any subscriber unit. These spatial signatures include the effects of the electronic signal paths in the base station hardware and any differences between the uplink and downlink electronic signal paths for the transponder or subscriber unit. One use of such information is to determine separate calibrations for each subscriber unit when the RF propagation to and from the subscriber unit is different. Another use is for calibrating the base station, but rather than obtaining a single calibration vector using the base station and a single transponder, using several transponders to determine the single calibration vector.

In one embodiment, the single calibration vector is the average calibration vector. In another embodiment, it is the weighted average calibration vector. The weighting given to the estimate made using a particular subscriber unit will depend on

a measure of the quality of the signal received by that subscriber unit, so that estimates from subscriber units having better quality signals are weighed more in the weighted average. A method and apparatus for determining signal quality is disclosed in International Application No. WO99/40689, published August 12, 1999 of Yun.

In the architecture of Figures 1 and 2, the base station DSP generates a set of signals that are used for calibration. In one example, all antennas transmit different known calibration signals so that the channel from each transmit antenna to each receive antenna can be calculated. Generally, after subtracting out the components specific to the transponder's location, a receive calibration vector can then be estimated from the difference in phase and amplitude with frequency of the channels from one transmit antenna to each receive antenna. By averaging the results from all the transmit antennas, the calibration vector can be improved still further. Correspondingly, a calibration vector of the transmit chains can be estimated, after subtracting out the transponder specific components, from the relative phases and amplitudes of the channels from different transmit antennas to one of the receive antennas. Again, averaging the results from all the different receive antennas can improve the estimate.

Using the two or more narrow band transponder returns, the relative phase and amplitude of the transmit and receive chains can be calibrated at two frequencies within the base station downlink and uplink bands, respectively. The measurements can also be used for calibrating group delay and any other frequency dependent differences between the receive or transmit chains. Higher accuracy can be obtained if the two narrow frequency bands are placed some distance apart within the traffic bands. Higher accuracy can also be obtained by using more than two different frequencies. The best choice of calibration frequencies and numbers of different frequencies will depend on the bandwidth of the traffic bands and the desired accuracy.

Because a group delay can be regarded as equivalent to a phase ramp with a specific slope, the relative difference in group delay among the transmit and receive chains, respectively, can be calibrated using the phase measurements. This can be done by computing the slopes of the phase ramps based on the phase measurements at the two frequencies within the bands. Since there is an ambiguity in each phase measurement due to phase wrapping, the relative phase between the two measurement frequencies can only be determined to within a phase window of 360 degrees. As a result, any group delay changes and differences within the delay corresponding to a

phase shift of 360 degrees between the two measurement frequencies can be measured and compensated for.

The group delay can be determined directly from a phase calibration process. If the system is calibrating the various receive and transmit chains for phase and amplitude differences, the phase determinations from that process can be used to find the group delay. Group delay can also be determined using relative phase measurements that are calculated apart from any phase calibration process. The phase calibration will give a calibration vector with a calibration coefficient α_{ij} for each antenna i and frequency j. The actual phase φ_{ij} of an antenna i at frequency j can be expressed as $\varphi_{ij} = \alpha_{ij} + \delta_j$, where δ_j is an arbitrary unknown phase term that is common to all antennas at frequency j. The value of δ need not be known in order to calibrate the transmit or receive chain with respect to the other chains. Only the relative phases characterized by the α's is needed.

For group delay, the difference between different transmit or receive chains is used. For a single frequency j, this difference $\Delta\varphi_j$ between antenna i and i' can be expressed as $\Delta\varphi_j = \varphi_{ij} - \varphi_{i'j} = \alpha_{ij} + \delta_j - (\alpha_{i'j} + \delta_j) = \alpha_{ij} - \alpha_{i'j}$. The group delay between the antennas i and i' is obtained by comparing the difference in phase $\Delta\varphi$ at different frequencies. For frequencies j and j', the group delay is therefore proportional to $\Delta\varphi_j - \Delta\varphi_{j'}$. Using the phase calibration vectors α's at the two different frequencies, the relative group delay can quickly be determined.

In the process described above, δ_j the arbitrary unknown phase term that is common to all antennas at frequency j remains unknown. This term can also vary over time. For example if frequency f_1 is repeatedly measured, the measured signature can be expressed as $e^{j\varphi}\mathbf{a}_1$, where $\mathbf{a}$ is the measurement vector at frequency f_1 containing elements $a_1, a_2, a_3, \ldots$ and the phase φ changes with each measurement. Alternatively, the measured phase can be normalized so that some component, for example, the first component, is real. in either case, the absolute phase is not measured.

As a result, the absolute group delay cannot easily be determined using the phase calibration values, however correcting for relative phase delays between the different transmit and receive chains significantly enhances performance. These relative phase differences constitute the differential phase delay between the transmit and receive chains of the system. Current digital signal processing technology can accommodate a frequency dependent phase variation from a single transmitter. If the phase variations from multiple transmitters can be aligned, then the variations in the

multiple transmitter system can be accommodated by the receiver in the same way as from a single transmitter. If the phase variations differ among the transmitters, the transmitted signal becomes much more difficult to resolve. Accordingly while a calibration that corrects for absolute group delay may be desirable in some applications, calibration for relative group delay is very useful. The more the differences between the transmit or alternatively, receive chains, can be reduced the higher the system's performance.

Using phase and amplitude measurements, calibration vectors can be formed and applied to transmissions by the base station. One approach uses spatial signatures from the receive chains of an antenna system and, using signatures at two different frequencies imposes a linear phase shift ramp. The spatial signatures can be made up of a vector or a set **a** of phase and amplitude measurements for each receive or transmit chain. They can be represented as $\mathbf{a}_j$ and $\mathbf{a}_{j'}$, where $\mathbf{a}_j$, for example, represents a set of values a_{j1}, a_{j2}, a_{j3}, ... a_M for each of M receive or transmit chains i = 1, 2, 3, ... M, at the frequency j. These two signatures are combined to derive the frequency dependent calibration factor $\mathbf{c}(f)$.

While a linear fit for $\mathbf{c}(f)$ provides for a simple and quick determination of the calibration vector using only two measured frequencies, as shown below, more frequencies can be measured and any variety of other curves or shapes can be matched to the measured results. The choice of an interpolation or curve matching algorithm as well as the choice of the number of different frequencies to measure will depend on a balance between calibration complexity and signal quality. The quality of the equalizers and the demodulators as well as the width of the frequency bandwidth of the system will likely also be considered among other factors.

To calibrate differential amplitude shifts with frequency, a frequency dependent amplitude calibration factor $| g_i(f) |$ for each antenna i = 1, ... M can be determined by linear interpolation:

$$| g_i(f) | = [(f-f_1)/(f_2-f_1)|a_{1,i}|] + [(f_2-f)/(f_2-f_1)|a_{2,i}|]$$

for $f_2 \geq f \geq f_1$, where f_2 corresponds to frequency j', f_1 corresponds to frequency j, $a_{1,i}$ corresponds to the phase and amplitude measurement for antenna i at frequency f_1 and $a_{2,i}$ corresponds to the phase amplitude measurement for antenna i at frequency f_2. Linear extrapolation can be used to extend the amplitude calibration factor outside the interval between the two measured frequencies f_1, f_2.

To determine a phase portion of the calibration vector $\mathbf{c}(f)$, a modified linear interpolation that compensates for the phase wrapping can be used. As mentioned

above, there is a relative phase window of 360 degrees or 2π, at which point, the phase wraps back around to zero. If angle (a) is an angle in degrees that can take any value from -180 degrees up to but not including 180, angle (a) $\in$ (-180, 180], and angle (a) corresponds to the complex number a, and a* is the complex conjugate of a, then the calibration phase $\varphi_i(f)$ for antenna i at frequency f can be expressed as shown below.

$$\varphi_i(f) = [\ (f-f_1)/(f_2-f_1)\ ]\ \text{angle}\ (a_{1,i})^* a_{2,i} + \text{angle}\ (a_{1,i})$$

for i = 1, ... M and the overall calibration factor is equal to the combination of the amplitude and phase calibration factors which can be expressed as shown below:

$$c_i(f) = |\ g_i(f)\ | e^{j(180/\pi)\varphi i(f)}$$

Method of Operation

An example of an operational process for calibrating a group of receive chains for group delay is shown in Figure 3. Other frequency dependent calibration vectors can be determined using a similar process. The calibration process typically includes calibrating the receive chain and the transmit chain with the same set of samples. Calibration of the transmit chains is shown in Figure 4. To begin a calibration cycle for the receive chain, the base station (BS) (see e.g. Figure 1) will generate a calibration signal. As discussed above, this is typically a narrowband signal at two or more frequencies. This narrowband transmit calibration signal is then transmitted from a single transmit chain of the base station 311. The transmission can occur at any time during the regular use of the base station for normal operation due to the small amount of additional energy added to the existing wideband data traffic by the narrowband signal. While only one transmit chain is required, transmitting from all of the transmit chains at once provides more samples for the receive calibration algorithms.

The transmitted narrowband calibration signal is received at the transponder 313, (see e.g. Figure 2). If the calibration signal is a wideband signal, it is converted to a set of at least two narrowband waveforms using appropriate bandpass filters as discussed above. If the signal has a particular spreading sequence or is modulated with a particular data or training sequence, this can be demodulated and a new signal can be modulated onto the signal. In one embodiment, the calibration signal is a narrowband signal, which is simply received, shifted in frequency 315, and transmitted back to the base station 317. This approach simplifies the transponder and

eliminates many other potential causes of errors. The frequency shifted calibration signal can also be shifted to two or more different frequencies and retransmitted so that calibration can be performed across different narrow frequency bands. However, the same effect can be achieved with a simpler transponder by sending several different calibration signals from the base station, each at a different frequency for the downlink. Each signal will be shifted to a different frequency for the uplink.

The base station receives the transponder signal at each of its receive antenna chains 319. These received transponder signals are sampled for each receive antenna chain 321 and the samples can be used to measure any number of characteristics of the received signal. Each set of samples from each receive chain represents a different view of the same narrowband transponder signal. To enhance reception, the DSP 217 will typically use narrow bandpass filters to eliminate most of the data traffic signal energy and isolate the received transponder signal. The received transponder signal is used to calculate a set of phases, for example the α's discussed above and amplitudes 323 The calculation in support of group delay will typically be based on comparing the received transponder signal as it was received by each receive chain to each signal as received by each other receive chain. This is commonly done by measuring phases and amplitudes and using a covariance matrix, for example. As an alternative, the signal can be sampled at only two receive chains. This will allow the two selected chains to be calibrated against each other. By repeating the process for each possible combination or for each receive chain against a receive chain selected to be the reference, a set of relative phase measurements can be obtained.

The process of transmitting and receiving calibration signals described above can then be repeated and the results averaged or stored 325. Further relative phases and amplitudes are calculated using the additional data 327 and a group delay is calculated 328. This group delay is typically in the form of a calibration vector composed of a set of phase and amplitude correction factors for each transmit and receive chain, as discussed above. Alternatively, the resulting calibration vector can be applied and the process repeated to find a new vector that is used to adjust the first vector. By applying the adjusted calibration vector after each cycle, the calibration should become progressively more accurate until it converges on the limit of the calibration system's accuracy. The transmission, reception and computations can be repeated for different combinations of receive chains and even for different transponders. Over time, the characteristics of the receive chains can change and so the process can also be repeated in order to update the calibration vectors with

changing conditions. When the calibrations are done against a reference chain, pairing each receive chain against the reference, the reference chain's vectors can be set at one, or some other normalized set of values, so that the vectors for the other receive chains represent the variance from the reference chain. Alternatively, the vectors can represent the variance from any other value, for example an average, mean or median response.

Calibration of the transmit chain is done in a similar way as shown in Figure 4. As with the receive chain, a calibration signal is transmitted to the transponder. In this case, the calibration signal is transmitted from each of the base station's transmit chains 329. So that they can be distinguished from each other when received, each receive chain uses a different modulation sequence. As with the receive calibration, this signal is a narrowband signal at at least two different frequencies. The narrowband signal allows the transponder to have a simple construction.

The calibration signals are received at the transponder 331. Which then, as with the receive calibration, shifts the frequency of the received calibration signals 333. After that, the shifted calibration signals are transmitted back to the base station 335. It is again possible to change modulated sequences or spreading codes but the simplest transponder will take the narrowband signal that it receives in the downlink band and transmit it back as a virtually identical narrowband signal in the uplink band.

The base station receives the transponder signals this time at just one receive antenna chain 337. The received transponder signals are sampled 339 and then the unique modulated sequences are used to extract each transmit chain calibration signal 341 from the sampled waveform. As with the receive calibration, a narrow bandpass filter is typically used to isolate the transponder signal. For calibration purposes, the transmitted calibration signals from each transmit chain are compared to each other 343. In order to make it easier to distinguish the simultaneously received signals from the different transmit chains, the number of simultaneous transmit chains can be reduced. For example, one of the transmit chains can be designated as the reference and then each other transmit chain can transmit with the reference, one pair at a time, until all the transmit chains have been calibrated against the reference. This is similar to the pair-wise receive chain calibration mentioned above.

These comparisons become the basis for generating a set of relative phases and amplitudes 345. The process of sending and receiving calibration signals can then be repeated 347 and further relative phases and amplitudes computed 349 to

refine the results. Then, the transmit group delay calibration vector can be calculated for each transmit chain 351. In one embodiment, the calibration vector determined in the first round is applied to each transmit chain, and then the process is repeated. The next calibration cycle will lead to greater accuracy as the gross errors have already been compensated. This is similar to performing a coarse tuning process and then a fine-tuning process.

The present invention provides many advantages over the prior art. Calibrations can be performed using only a simple, inexpensive transponder. Both transmit and receive calibration can be determined in a single transaction and the method self-corrects for reference frequency offsets in the antenna array system. Accordingly, calibration in accordance with the present invention is inherently accurate. While the invention has been described primarily as a calibration of a base station using a remote transponder, it can be applied to remote user terminals that have multiple antennas. It can also be applied to any other type of wireless network with multiple antenna system whether one with base stations and remotes, equal peers or masters and slaves.

To improve the reception of regular traffic during calibration, it may be desirable to apply a notch filter at the base station to filter out the transponder signal bands. This would typically be a digital filter and can be turned off when no calibration signal is active. The subscriber units could similarly have a notch filter for the calibration signal from the base station.

In the description above, for the purposes of explanation, numerous specific details are set forth in order to provide a thorough understanding of the present invention. It will be apparent, however, to one skilled in the art that the present invention may be practiced without some of these specific details. In other instances, well-known structures and devices are shown in block diagram form.

The present invention includes various steps. The steps of the present invention may be performed by hardware components, such as those shown in Figures 1 and 2, or may be embodied in machine-executable instructions, which may be used to cause a general-purpose or special-purpose processor or logic circuits, such as a DSP programmed with the instructions to perform the steps. Alternatively, the steps may be performed by a combination of hardware and software.

The present invention may be provided as a computer program product which may include a machine-readable medium having stored thereon instructions which may be used to program a computer (or other electronic devices) to perform a process

according to the present invention. The machine-readable medium may include, but is not limited to, floppy diskettes, optical disks, CD-ROMs, and magneto-optical disks, ROMs, RAMs, EPROMs, EEPROMs, magnet or optical cards, flash memory, or other type of media or machine-readable medium suitable for storing electronic instructions. Moreover, the present invention may also be downloaded as a computer program product, wherein the program may be transferred from a remote computer to a requesting computer by way of data signals embodied in a carrier wave or other propagation medium via a communication link (e.g., a modem or network connection).

Importantly, while the present invention has been described in the context of a wireless spread spectrum data system for mobile remote terminals, it can be applied to a wide variety of different wireless systems in which data is exchanged. Such systems include voice, video, music, broadcast and other types of data systems without external connections. The present invention can be applied to fixed user terminals as well as to low and high mobility terminals. Many of the methods are described herein in a basic form but steps can be added to or deleted from any of the methods and information can be added or subtracted from any of the described messages without departing from the basic scope of the present invention. It will be apparent to those skilled in the art that many further modifications and adaptations can be made. The particular embodiments are not provided to limit the invention but to illustrate it. The scope of the present invention is not to be determined by the specific examples provided above but only by the claims below.

CLAIMS

What is claimed is:

1. A radio communications system comprising:

an antenna array adapted to transmit and receive radio communications signals with a plurality of other terminals, the communications signals each using a particular minimum bandwidth;

a transmit chain to transmit a calibration signal through the antenna array to a transponder on at least two different frequency bands within the minimum bandwidth;

a receive chain to receive through the antenna array a transponder signal from the transponder, the transponder signal being received on at least two different frequency bands and being based on the calibration signal; and

a signal processor to determine a frequency dependent calibration vector based on the at least two frequency bands of the transponder signal as received through the receive chain.

2. The system of claim 1, wherein determining a frequency dependent calibration vector comprises comparing relative phases for the transponder signal at a first one of the at least two frequencies to relative phases for the transponder signal at a second one of the at least two frequencies to determine a group delay.

3. The system of claim 1, wherein the transponder signal is shifted in frequency as compared to the calibration signal.

4. The system of claim 1, further comprising measuring the relative phases and amplitudes of the transponder signal as received by the receive chain.

5. The system of claim 4:

wherein the receive chain comprises a plurality of receive chains;

wherein each receive chain receives the transponder signal; and

wherein the signal processor determines a group delay by comparing the relative phases of the transponder signal at each frequency as received by each receive chain.

6. The system of claim 5 wherein determining a frequency dependent calibration vector comprises determining a receive chain group delay by comparing a phase difference between at least two receive chains for the transponder signal at a first one of the at least two frequency bands to a phase difference between the same two receive chains for the transponder signal at a second one of the at least two frequency bands.

7. The system of claim 6, wherein one of the plurality of receive chains is selected as a reference receive chain and the group delay for each receive chain is characterized with respect to the reference receive chain.

8. The system of Claim 4 wherein the signal processor determines an uplink signature of the transponder at the antenna array at each frequency of the transponder signal using measured phases and amplitudes of the transponder signal and wherein the signal processor determines the frequency dependent calibration vector for the receive chain using the uplink signatures of the transponder.

9. The system of Claim 4 wherein the signal processor determines a downlink signature of the transmit chain at the transponder using measured phases and amplitudes at each frequency of the transponder signal and wherein the signal processor further determines the frequency dependent calibration vector for the transmit chain using the downlink signatures of the transmit chain.

10. The system of claim 1:

wherein the transmit chain comprises a plurality of transmit chains;

wherein each transmit chain transmits the calibration signal; and

wherein the signal processor determines a frequency dependent transmit calibration vector by comparing the relative phases of the transponder signal at each frequency of the transponder signal as received by each receive chain.

11. The system of claim 10, wherein the calibration signal comprises a plurality of signals, one from each transmit chain, each signal being individually identifiable based on a unique modulation sequence.

12. The system of claim 10 wherein determining a frequency dependent transmit calibration vector comprises comparing a phase difference between two transmit chains for the transponder signal at a first one of the at least two frequencies to a phase difference between the same two transmit chains for the transponder signal at a second one of the at least two frequencies to determine a group delay.

13. The system of Claim 12 wherein one of the plurality of transmit chains is selected as a reference chain and the group delay of each transmit chain is defined with respect to the reference chain.

14. A machine-readable medium having stored thereon data representing instructions which, when executed by a machine, cause the machine to perform operations comprising:

transmitting radio communications signals to a plurality of other terminals using a transmit chain, the communications signals each using a particular minimum transmit bandwidth;

receiving radio communications signals from a plurality of other terminals using a receive chain, the communications signals each using a particular minimum receive bandwidth;

transmitting a calibration signal through the transmit chain to a transponder on at least two different frequency bands within the minimum transmit bandwidth;

receiving a transponder signal through the receive chain from the transponder, the transponder signal being received on at least two different frequency bands within the minimum receive bandwidth and being based on the calibration signal; and

determining a frequency dependent calibration vector based on the at least two frequency bands of the transponder signal as received through the receive chain.

15. The medium of claim 14, wherein determining a frequency dependent calibration vector comprises comparing relative phases for the transponder signal at a first one of the at least two frequencies to relative phases for the transponder signal at a second one of the at least two frequencies to determine a group delay.

16. The medium of claim 14, wherein the transponder signal is shifted in frequency as compared to the calibration signal.

17. The medium of claim 14 wherein determining a frequency dependent calibration vector comprises determining a receive chain group delay by comparing a phase difference between at least two receive chains for the transponder signal at a first one of the at least two frequency bands to a phase difference between the same two receive chains for the transponder signal at a second one of the at least two frequency bands.

18. The medium of claim 14 wherein determining a frequency dependent transmit calibration vector comprises comparing a phase difference between two transmit chains for the transponder signal at a first one of the at least two frequencies to a phase difference between the same two transmit chains for the transponder signal at a second one of the at least two frequencies to determine a group delay.

19. A method comprising:

transmitting radio communications signals to a plurality of other terminals using a transmit chain, the communications signals each using a particular minimum transmit bandwidth;

receiving radio communications signals from a plurality of other terminals using a receive chain, the communications signals each using a particular minimum receive bandwidth;

transmitting a calibration signal through the transmit chain to a transponder on at least two different frequency bands within the minimum transmit bandwidth;

receiving a transponder signal through the receive chain from the transponder, the transponder signal being received on at least two different frequency bands within the minimum receive bandwidth and being based on the calibration signal; and

determining a frequency dependent calibration vector based on the at least two frequency bands of the transponder signal as received through the receive chain.

20. The method of claim 19, wherein determining a frequency dependent calibration vector comprises measuring the relative phases and amplitudes of the transponder signal as received by the receive chain.

21. The method of claim 19, wherein determining a frequency dependent calibration vector comprises determining a group delay by comparing the relative phases of the transponder signal at each frequency as received by a plurality of receive chains.

22. The method of claim 21, wherein one of the plurality of receive chains is selected as a reference receive chain and the group delay for each receive chain is characterized with respect to the reference receive chain.

23. The method of Claim 21 wherein determining a frequency dependent calibration vector comprises determining an uplink signature of the transponder at the receive chains at each frequency of the transponder signal using measured phases and amplitudes of the transponder signal and determining the frequency dependent calibration vector for the receive chains using the uplink signatures of the transponder.

24. The method of Claim 21 wherein the signal processor determining a frequency dependent calibration vector comprises determining a downlink signature of a plurality of transmit chains at the transponder using measured phases and amplitudes at each frequency of the transponder signal and determining the frequency dependent calibration vector for the transmit chains using the downlink signatures of the transmit chains.

25. The method of claim 19 wherein determining a frequency dependent calibration vector comprises determining a frequency dependent transmit calibration

vector by comparing the relative phases of the transponder signal at each frequency of the transponder signal as received by each of a plurality of receive chains.

26. The system of claim 1, wherein the system is a code division multiple access system.

27. The medium of claim 14, wherein the radio communications signals conform to a standard for code division multiple access.

28. The method of claim 19, wherein the radio communications signals conform to a standard for code division multiple access.

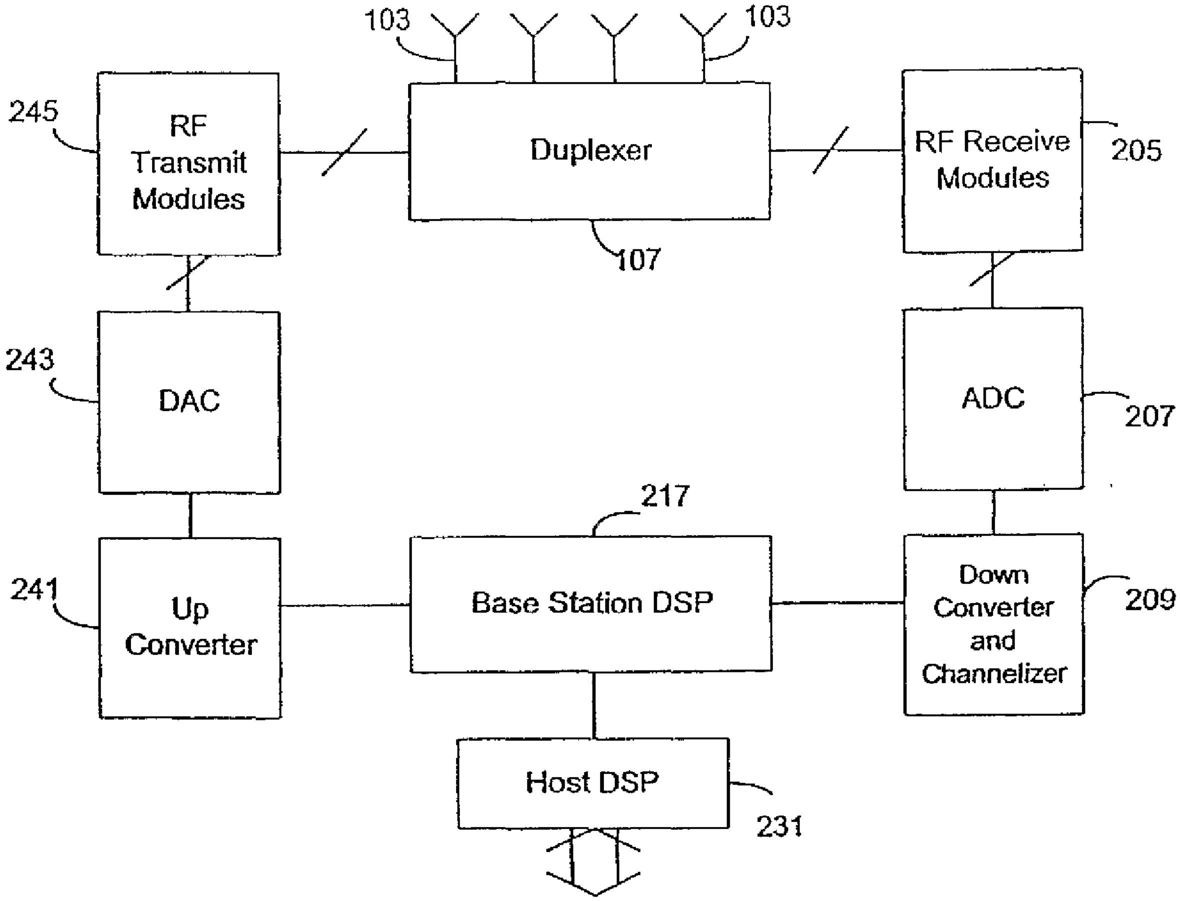
103
103
245
RF Transmit Modules
Duplexer
RF Receive Modules
205
107
243
DAC
ADC
207
217
241
Up Converter
Base Station DSP
Down Converter and Channelizer
209
Host DSP
231

Figure 1

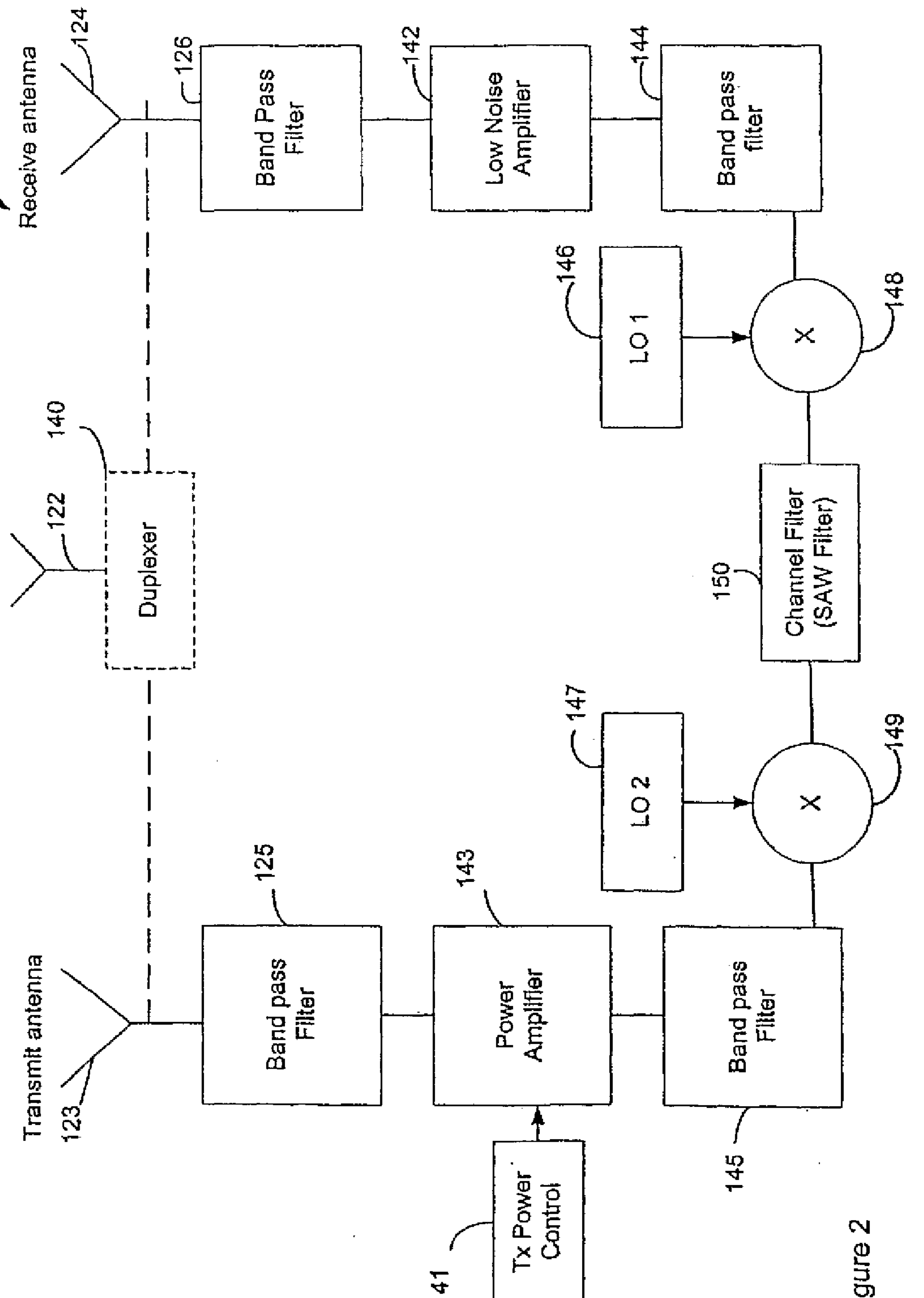
118
Receive antenna
124
126
Band Pass Filter
142
Low Noise Amplifier
144
Band pass filter
146
LO 1
148
X
140
122
Duplexer
150
Channel Filter (SAW Filter)
147
LO 2
149
X
Transmit antenna
123
125
Band pass Filter
143
Power Amplifier
Band pass Filter
145
141
Tx Power Control

Figure 2

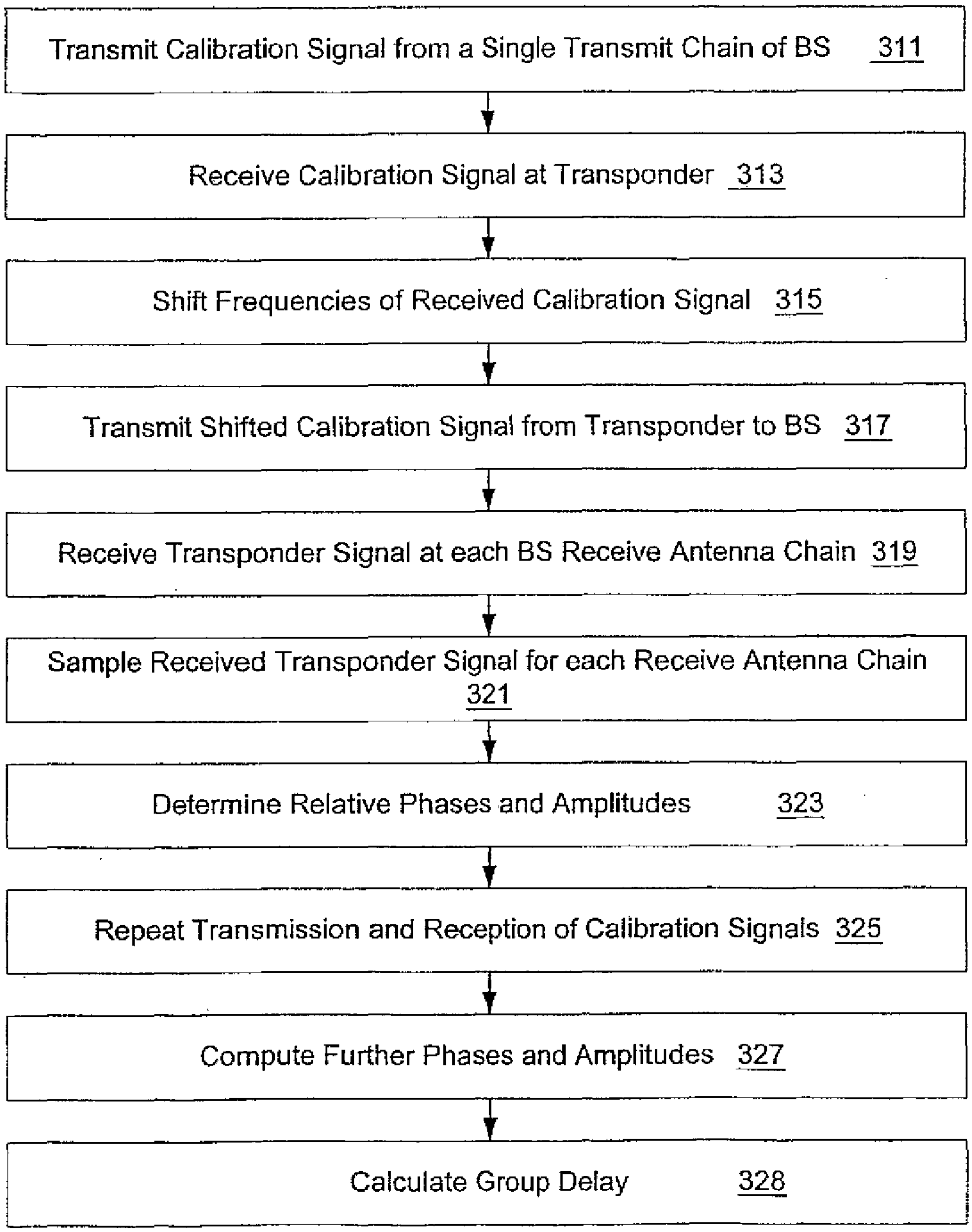
Transmit Calibration Signal from a Single Transmit Chain of BS 311
Receive Calibration Signal at Transponder 313
Shift Frequencies of Received Calibration Signal 315
Transmit Shifted Calibration Signal from Transponder to BS 317
Receive Transponder Signal at each BS Receive Antenna Chain 319
Sample Received Transponder Signal for each Receive Antenna Chain 321
Determine Relative Phases and Amplitudes 323
Repeat Transmission and Reception of Calibration Signals 325
Compute Further Phases and Amplitudes 327
Calculate Group Delay 328

Figure 3

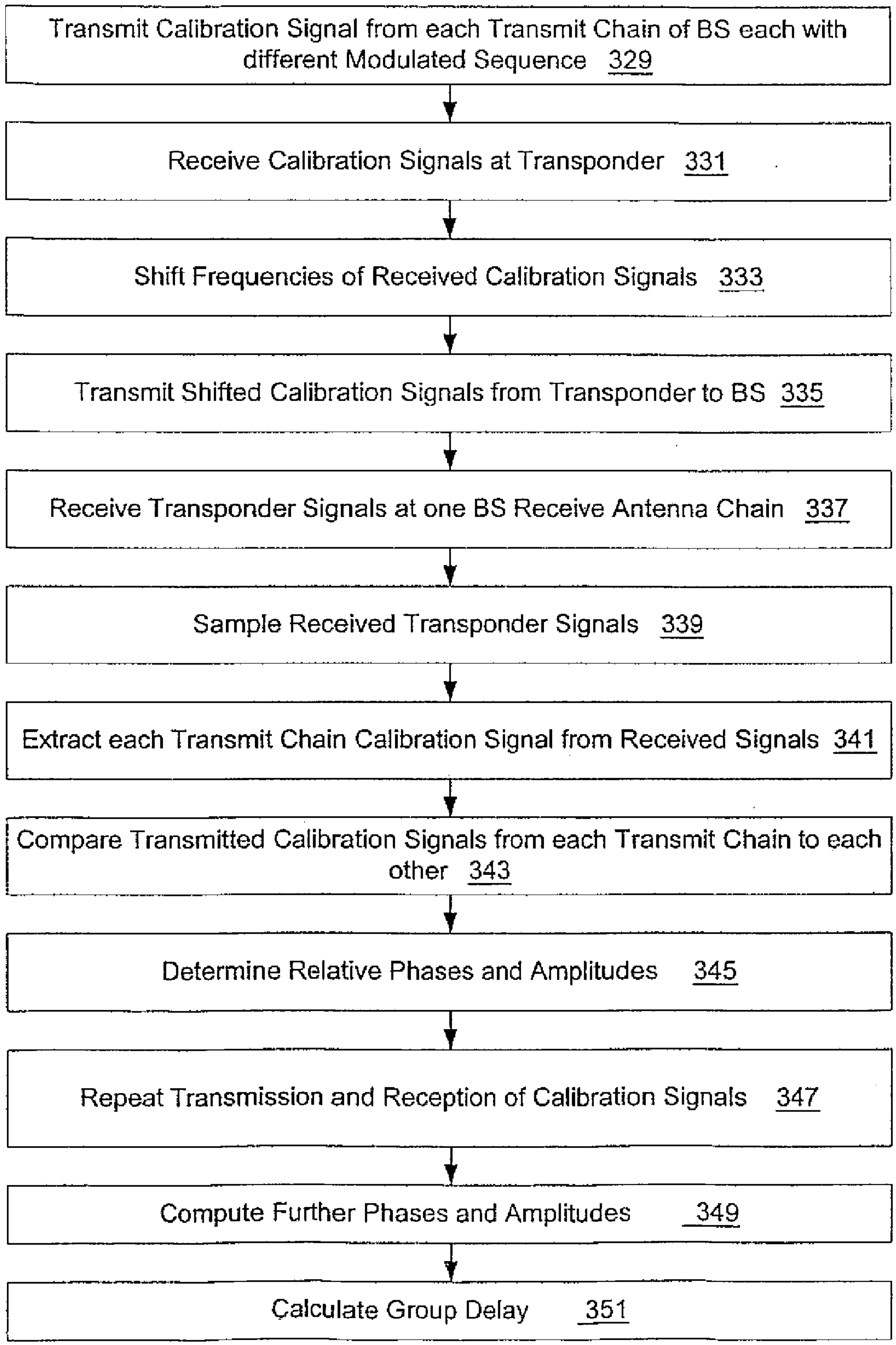
Transmit Calibration Signal from each Transmit Chain of BS each with different Modulated Sequence 329
Receive Calibration Signals at Transponder 331
Shift Frequencies of Received Calibration Signals 333
Transmit Shifted Calibration Signals from Transponder to BS 335
Receive Transponder Signals at one BS Receive Antenna Chain 337
Sample Received Transponder Signals 339
Extract each Transmit Chain Calibration Signal from Received Signals 341
Compare Transmitted Calibration Signals from each Transmit Chain to each other 343
Determine Relative Phases and Amplitudes 345
Repeat Transmission and Reception of Calibration Signals 347
Compute Further Phases and Amplitudes 349
Calculate Group Delay 351

Figure 4

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 02/30896

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H01Q3/26

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H01Q

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data, INSPEC, COMPENDEX

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US 5 546 090 A (ROY III RICHARD H ET AL) 13 August 1996 (1996-08-13) cited in the application abstract column 2, line 13-39 column 4, line 20 -column 8, line 53 figures 4-7 ---	1,3,4, 14,16, 19,20, 26-28
X	US 5 294 934 A (MATSUMOTO SOICHI) 15 March 1994 (1994-03-15) abstract column 4, line 63 -column 6, line 7 figure 1 --- -/--	1,14,19

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
11 December 2002	20/12/2002

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Helms, J

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 02/30896

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US 6 236 839 B1 (GU YUCONG ET AL) 22 May 2001 (2001-05-22) abstract column 1, line 37 -column 4, line 8 figure 1	1,14,19
A	US 6 037 898 A (BARRATT CRAIG H ET AL) 14 March 2000 (2000-03-14) cited in the application abstract column 9, line 49 -column 18, line 65 figure 2	1-28

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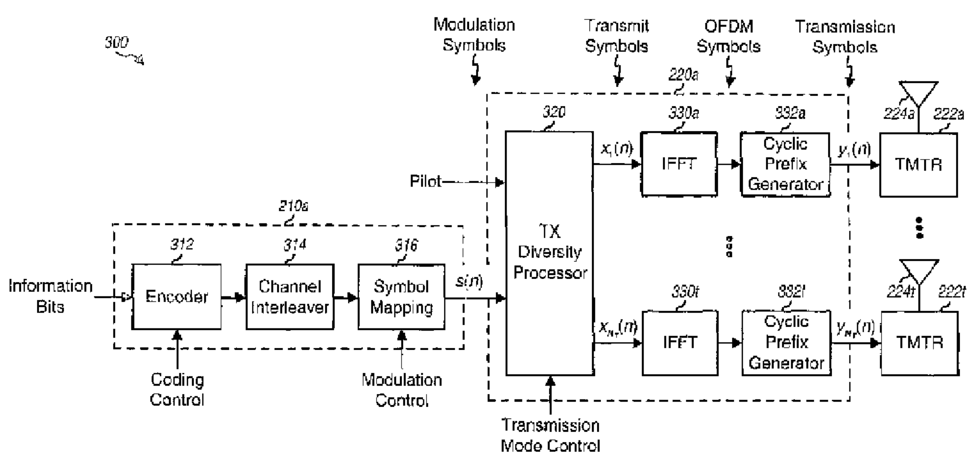
(54) Title: DIVERSITY TRANSMISSION MODES FOR MIMO OFDM COMMUNICATION SYSTEMS



(57) Abstract: Techniques for transmitting data using a number of diversity transmission modes to improve reliability. At a transmitter, for each of one or more data streams, a particular diversity transmission mode is selected for use from among a number of possible transmission modes. These transmission modes may include a frequency diversity transmission mode, a Walsh diversity transmission mode, a space time transmit diversity (STTD) transmission mode, and a Walsh-STTD transmission mode. Each diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof. Each data stream is coded and modulated to provide modulation symbols, which are further processed based on the selected diversity transmission mode to provide transmit symbols. For OFDM, the transmit symbols for all data streams are further OFDM modulated to provide a stream of transmission symbols for each transmit antenna used for data transmission.

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DIVERSITY TRANSMISSION MODES FOR MIMO OFDM COMMUNICATION SYSTEMS

BACKGROUND

Field

[1001] The present invention relates generally to data communication, and more specifically to techniques for transmitting data using a number of diversity transmission modes in MIMO OFDM systems.

Background

[1002] Wireless communication systems are widely deployed to provide various types of communication such as voice, packet data, and so on. These systems may be multiple-access systems capable of supporting communication with multiple users either sequentially or simultaneously. This is achieved by sharing the available system resources, which may be quantified by the total available operating bandwidth and transmit power.

[1003] A multiple-access system may include a number of access points (or base stations) that communicate with a number of user terminals. Each access point may be equipped with one or multiple antennas for transmitting and receiving data. Similarly, each terminal may be equipped with one or multiple antennas.

[1004] The transmission between a given access point and a given terminal may be characterized by the number of antennas used for data transmission and reception. In particular, the access point and terminal pair may be viewed as (1) a multiple-input multiple-output (MIMO) system if multiple (N_T) transmit antennas and multiple (N_R) receive antennas are employed for data transmission, (2) a multiple-input single-output (MISO) system if multiple transmit antennas and a single receive antenna are employed, (3) a single-input multiple-output (SIMO) system if a single transmit antenna and multiple receive antennas are employed, or (4) a single-input single-output (SISO) system if a single transmit antenna and a single receive antenna are employed.

[1005] For a MIMO system, a MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S independent channels is also referred to as a spatial

subchannel of the MIMO channel and corresponds to a dimension. The MIMO system can provide improved performance (e.g., increased transmission capacity and/or greater reliability) if the additional dimensionalities created by the multiple transmit and receive antennas are utilized. For a MISO system, only one spatial subchannel is available for data transmission. However, the multiple transmit antennas may be used to transmit data in a manner to improve the likelihood of correct reception by the receiver.

[1006] The spatial subchannels of a wideband system may encounter different channel conditions due to various factors such as fading and multipath. Each spatial subchannel may thus experience frequency selective fading, which is characterized by different channel gains at different frequencies of the overall system bandwidth. It is well known that frequency selective fading causes inter-symbol interference (ISI), which is a phenomenon whereby each symbol in a received signal acts as distortion to subsequent symbols in the received signal. The ISI distortion degrades performance by impacting the ability to correctly detect the received symbols.

[1007] To combat frequency selective fading, orthogonal frequency division multiplexing (OFDM) may be used to effectively partition the overall system bandwidth into a number of (N_F) subbands, which may also be referred to as OFDM subbands, frequency bins, or frequency sub-channels. Each subband is associated with a respective subcarrier upon which data may be modulated. For each time interval that may be dependent on the bandwidth of one subband, a modulation symbol may be transmitted on each of the N_F subbands.

[1008] For a multiple-access system, a given access point may communicate with terminals having different number of antennas at different times. Moreover, the characteristics of the communication channels between the access point and the terminals typically vary from terminal to terminal and may further vary over time, especially for mobile terminals. Different transmission schemes may then be needed for different terminals depending on their capabilities and requirements.

[1009] There is therefore a need in the art for techniques for transmitting data using a number of diversity transmission modes depending on the capability of the receiver device and the channel conditions.

SUMMARY

[1010] Techniques are provided herein for transmitting data in a manner to improve the reliability of data transmission. A MIMO OFDM system may be designed to support a number of modes of operation for data transmission. These transmission modes may include diversity transmission modes, which may be used to achieve higher reliability for certain data transmission (e.g., for overhead channels, poor channel conditions, and so on). The diversity transmission modes attempt to achieve transmit diversity by establishing orthogonality among multiple signals transmitted from multiple transmit antennas. Orthogonality among the transmitted signals may be attained in frequency, time, space, or any combination thereof. The transmission modes may also include spatial multiplexing transmission modes and beam steering transmission modes, which may be used to achieve higher bit rates under certain favorable channel conditions.

[1011] In an embodiment, a method is provided for processing data for transmission in a wireless (e.g., MIMO OFDM) communication system. In accordance with the method, a particular diversity transmission mode to use for each of one or more data streams is selected from among a number of possible transmission modes. Each diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof. Each data stream is coded and modulated based on coding and modulation schemes selected for the data stream to provide modulation symbols. The modulation symbols for each data stream are further processed based on the selected diversity transmission mode to provide transmit symbols. For OFDM, the transmit symbols for all data streams are further OFDM modulated to provide a stream of transmission symbols for each of one or more transmit antennas used for data transmission. Pilot symbols may also be multiplexed with the modulation symbols using frequency division multiplexing (FDM), time division multiplexing (TDM), code division multiplexing (CDM), or any combination thereof.

[1012] The transmission modes may include, for example, (1) a frequency diversity transmission mode that redundantly transmits modulation symbols over multiple OFDM subbands, (2) a Walsh diversity transmission mode that transmits each modulation symbol over N_T OFDM symbol periods, where N_T is the number of transmit antennas used for data transmission, (3) a space time transmit diversity (STTD) transmission

mode that transmits modulation symbols over multiple OFDM symbol periods and multiple transmit antennas, and (4) a Walsh-STTD transmission mode that transmits modulation symbols using a combination of Walsh diversity and STTD. For the Walsh diversity and Walsh-STTD transmission modes, the same modulation symbols may be redundantly transmitted over all transmit antennas or different modulation symbols may be transmitted over different transmit antennas.

[1013] Each data stream may be for an overhead channel or targeted for a specific receiver device. The data rate for each user-specific data stream may be adjusted based on the transmission capability of the receiver device. The transmit symbols for each data stream are transmitted on a respective group of one or more subbands.

[1014] In another embodiment, a method is provided for processing a data transmission at a receiver of a wireless communication system. In accordance with the method, the particular diversity transmission mode used for each of one or more data streams to be recovered is initially determined. The diversity transmission mode used for each is selected from among a number of possible transmission modes. Received symbols for each data stream are then processed based on the diversity transmission mode used for the data stream to provide recovered symbols, which are estimates of modulation symbols transmitted from a transmitter for the data stream. The recovered symbols for each data stream are further demodulated and decoded to provide decoded data for the data stream.

[1015] Various aspects and embodiments of the invention are described in further detail below. The invention further provides methods, transmitter units, receiver units, terminals, access points, systems, and other apparatuses and elements that implement various aspects, embodiments, and features of the invention, as described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[1016] The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[1017] FIG. 1 is a diagram of a multiple-access system that supports a number of users;

[1018] FIG. 2 is a block diagram of an embodiment of an access point and two terminals;

[1019] FIG. 3 is a block diagram of a transmitter unit;

[1020] FIG. 4 is a block diagram of a TX diversity processor that may be used to implement the frequency diversity scheme;

[1021] FIG. 5 is a block diagram of a TX diversity processor that may be used to implement the Walsh diversity scheme;

[1022] FIG. 6 is a block diagram of a TX diversity processor that may be used to implement the STTD scheme;

[1023] FIG. 7 is a block diagram of a TX diversity processor that may be used to implement a repeated Walsh-STTD scheme;

[1024] FIG. 8 is a block diagram of a TX diversity processor that may be used to implement a non-repeated Walsh-STTD scheme;

[1025] FIG. 9 is a block diagram of a receiver unit;

[1026] FIG. 10 is a block diagram of an RX diversity processor;

[1027] FIG. 11 is a block diagram of an RX antenna processor within the RX diversity processor and which may be used for the Walsh diversity scheme; and

[1028] FIG. 12 is a block diagram of an RX subband processor within the RX antenna processor and which may be used for the repeated and non-repeated Walsh-STTD schemes.

DETAILED DESCRIPTION

[1029] **FIG. 1** is a diagram of a multiple-access system 100 that supports a number of users. System 100 includes one or more access points (AP) 104 that communicate with a number of terminals (T) 106 (only one access point is shown in FIG. 1 for simplicity). An access point may also be referred to as a base station, a UTRAN, or some other terminology. A terminal may also be referred to as a handset, a mobile station, a remote station, user equipment (UE), or some other terminology. Each terminal 106 may concurrently communicate with multiple access points 104 when in soft handoff (if soft handoff is supported by the system).

[1030] In an embodiment, each access point 104 employs multiple antennas and represents (1) the multiple-input (MI) for a downlink transmission from the access point to a terminal and (2) the multiple-output (MO) for an uplink transmission from the terminal to the access point. A set of one or more terminals 106 communicating with a given access point collectively represents the multiple-output for the downlink transmission and the multiple-input for the uplink transmission.

[1031] Each access point 104 can communicate with one or multiple terminals 106, either concurrently or sequentially, via the multiple antennas available at the access point and the one or more antennas available at each terminal. Terminals not in active communication may receive pilots and/or other signaling information from the access point, as shown by the dashed lines for terminals 106e through 106h in FIG. 1.

[1032] For the downlink, the access point employs N_T antennas and each terminal employs 1 or N_R antennas for reception of one or more data streams from the access point. In general, N_R can be different for different multi-antenna terminals and can be any integer. A MIMO channel formed by the N_T transmit antennas and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each such independent channel is also referred to as a spatial subchannel of the MIMO channel. The terminals concurrently receiving downlink data transmission need not be equipped with equal number of receive antennas.

[1033] For the downlink, the number of receive antennas at a given terminal may be equal to or greater than the number of transmit antennas at the access point (i.e., $N_R \geq N_T$). For such a terminal, the number of spatial subchannels is limited by the number of transmit antennas at the access point. Each multi-antenna terminal communicates with the access point via a respective MIMO channel formed by the access point's N_T transmit antennas and its own N_R receive antennas. However, even if multiple multi-antenna terminals are selected for concurrent downlink data transmission, only N_S spatial subchannels are available regardless of the number of terminals receiving the downlink transmission.

[1034] For the downlink, the number of receive antennas at a given terminal may also be less than the number of transmit antennas at the access point (i.e., $N_R < N_T$). For example, a MISO terminal is equipped with a single receive antenna ($N_R = 1$) for downlink data transmission. The access point may then employ diversity, beam

steering, space division multiple access (SDMA), or some other transmission techniques to communicate simultaneously with one or multiple MISO terminals.

[1035] For the uplink, each terminal may employ a single antenna or multiple antennas for uplink data transmission. Each terminal may also utilize all or only a subset of its available antennas for uplink transmission. At any given moment, the N_T transmit antennas for the uplink are formed by all antennas used by one or more active terminals. The MIMO channel is then formed by the N_T transmit antennas from all active terminals and the access point's N_R receive antennas. The number of spatial subchannels is limited by the number of transmit antennas, which is typically limited by the number of receive antennas at the access point (i.e., $N_S \leq \min\{N_T, N_R\}$).

[1036] **FIG. 2** is a block diagram of an embodiment of access point 104 and two terminals 106. On the downlink, at access point 104, various types of traffic data such as user-specific data from a data source 208, signaling, and so on are provided to a transmit (TX) data processor 210. Processor 210 then formats and encodes the traffic data based on one or more coding schemes to provide coded data. The coded data is then interleaved and further modulated (i.e., symbol mapped) based on one or more modulation schemes to provide modulation symbols (i.e., modulated data). The data rate, coding, interleaving, and symbol mapping may be determined by controls provided by a controller 230 and a scheduler 234. The processing by TX data processor 210 is described in further detail below.

[1037] A transmit processor 220 then receives and processes the modulation symbols and pilot data to provide transmission symbols. The pilot data is typically known data processed in a known manner, if at all. In a specific embodiment, the processing by transmit processor 220 includes (1) processing the modulation symbols based on one or more transmission modes selected for use for data transmission to the terminals to provide transmit symbols and (2) OFDM processing the transmit symbols to provide transmission symbols. The processing by transmit processor 220 is described in further detail below.

[1038] Transmit processor 220 provides N_T streams of transmission symbols to N_T transmitters (TMTR) 222a through 222t, one transmitter for each antenna used for data transmission. Each transmitter 222 converts its transmission symbol stream into one or more analog signals and further conditions (e.g., amplifies, filters, and frequency

upconverts) the analog signals to generate a respective downlink modulated signal suitable for transmission over a wireless communication channel. Each downlink modulated signal is then transmitted via a respective antenna 224 to the terminals.

[1039] At each terminal 106, the downlink modulated signals from multiple transmit antennas of the access point are received by one or multiple antennas 252 available at the terminal. The received signal from each antenna 252 is provided to a respective receiver (RCVR) 254. Each receiver 254 conditions (e.g., filters, amplifies, and frequency downconverts) its received signal and further digitizes the conditioned signal to provide a respective stream of samples.

[1040] A receive processor 260 then receives and processes the streams of samples from all receivers 254 to provide recovered symbols (i.e., demodulated data). In a specific embodiment, the processing by receive processor 260 includes (1) OFDM processing the received transmission symbols to provide received symbols, and (2) processing the received symbols based on the selected transmission mode(s) to obtain recovered symbols. The recovered symbols are estimates of the modulation symbols transmitted by the access point. The processing by receive processor 260 is described in further detail below.

[1041] A receive (RX) data processor 262 then symbol demaps, deinterleaves, and decodes the recovered symbols to obtain the user-specific data and signaling transmitted on the downlink for the terminal. The processing by receive processor 260 and RX data processor 262 is complementary to that performed by transmit processor 220 and TX data processor 210, respectively, at the access point.

[1042] On the uplink, at terminal 106, various types of traffic data such as user-specific data from a data source 276, signaling, and so on are provided to a TX data processor 278. Processor 278 codes the different types of traffic data in accordance with their respective coding schemes to provide coded data and further interleaves the coded data. A modulator 280 then symbol maps the interleaved data to provide modulated data, which is provided to one or more transmitters 254. OFDM may or may not be used for the uplink data transmission, depending on the system design. Each transmitter 254 conditions the received modulated data to generate a respective uplink modulated signal, which is then transmitted via an associated antenna 252 to the access point.

[1043] At access point 104, the uplink modulated signals from one or more terminals are received by antennas 224. The received signal from each antenna 224 is provided to a receiver 222, which conditions and digitizes the received signal to provide a respective stream of samples. The sample streams from all receivers 222 are then processed by a demodulator 240 and further decoded (if necessary) by an RX data processor 242 to recover the data transmitted by the terminals.

[1044] Controllers 230 and 270 direct the operation at the access point and the terminal, respectively. Memories 232 and 272 provide storage for program codes and data used by controllers 230 and 270, respectively. Scheduler 234 schedules the data transmission on the downlink (and possibly the uplink) for the terminals.

[1045] For clarity, various transmit diversity schemes are specifically described below for downlink transmission. These schemes may also be used for uplink transmission, and this is within the scope of the invention. Also for clarity, in the following description, subscript "*i*" is used as an index for the receive antennas, subscript "*j*" is used as an index for the transmit antennas, and subscript "*k*" is used as an index for the subbands in the MIMO OFDM system.

Transmitter Unit

[1046] **FIG. 3** is a block diagram of a transmitter unit 300, which is an embodiment of the transmitter portion of access point 104. Transmitter unit 300 includes (1) a TX data processor 210a that receives and processes traffic and pilot data to provide modulation symbols and (2) a transmit processor 220a that further processes the modulation symbols to provide N_T streams of transmission symbols for the N_T transmit antennas. TX data processor 210a and transmit processor 220a are one embodiment of TX data processor 210 and transmit processor 220, respectively, in FIG. 2.

[1047] In the specific embodiment shown in FIG. 3, TX data processor 210a includes an encoder 312, a channel interleaver 314, and a symbol mapping element 316. Encoder 312 receives and codes the traffic data (i.e., the information bits) based on one or more coding schemes to provide coded bits. The coding increases the reliability of the data transmission.

[1048] In an embodiment, the user-specific data for each terminal and the data for each overhead channel may be considered as distinct data streams. The overhead

channels may include broadcast, paging, and other common channels intended to be received by all terminals. Multiple data streams may also be sent to a given terminal. Each data stream may be coded independently based on a specific coding scheme selected for that data stream. Thus, a number of independently coded data streams may be provided by encoder 312 for different overhead channels and terminals.

[1049] The specific coding scheme to be used for each data stream is determined by a coding control from controller 230. The coding scheme for each terminal may be selected, for example, based on feedback information received from the terminal. Each coding scheme may include any combination of forward error detection (FED) codes (e.g., a cyclic redundancy check (CRC) code) and forward error correction (FEC) codes (e.g., a convolutional code, a Turbo code, a block code, and so on). A coding scheme may also designate no coding at all. Binary or trellis-based codes may also be used for each data stream. Moreover, with convolutional and Turbo codes, puncturing may be used to adjust the code rate. More specifically, puncturing may be used to increase the code rate above the base code rate.

[1050] In a specific embodiment, the data for each data stream is initially partitioned into frames (or packets). For each frame, the data may be used to generate a set of CRC bits for the frame, which is then appended to the data. The data and CRC bits for each frame are then coded with either a convolutional code or a Turbo code to generate the coded data for the frame.

[1051] Channel interleaver 314 receives and interleaves the coded bits based on one or more interleaving schemes. Typically, each coding scheme is associated with a corresponding interleaving scheme. In this case, each independently coded data stream would be interleaved separately. The interleaving provides time diversity for the coded bits, permits each data stream to be transmitted based on an average SNR of the subbands and spatial subchannels used for the data stream, combats fading, and further removes correlation between coded bits used to form each modulation symbol.

[1052] With OFDM, the channel interleaver may be designed to distribute the coded data for each data stream over multiple subbands of a single OFDM symbol or possibly over multiple OFDM symbols. The objective of the channel interleaver is to randomize the coded data so that the likelihood of consecutive coded bits being corrupted by the communication channel is reduced. When the interleaving interval for a given data

stream spans a single OFDM symbol, the coded bits for the data stream are randomly distributed across the subbands used for the data stream to exploit frequency diversity. When the interleaving interval spans multiple OFDM symbols, the coded bits are randomly distributed across the data-carrying subbands and the multi-symbol interleaving interval to exploit both frequency and time diversity. For a wireless local area network (WLAN), the time diversity realized by interleaving over multiple OFDM symbols may not be significant if the minimum expected coherence time of the communication channel is many times longer than the interleaving interval.

[1053] Symbol mapping element 316 receives and maps the interleaved data in accordance with one or more modulation schemes to provide modulation symbols. A particular modulation scheme may be used for each data stream. The symbol mapping for each data stream may be achieved by grouping sets of q_m coded and interleaved bits to form data symbols (each of which may be a non-binary value), and mapping each data symbol to a point in a signal constellation corresponding to the modulation scheme selected for use for that data stream. The selected modulation scheme may be QPSK, M-PSK, M-QAM, or some other modulation scheme. Each mapped signal point is a complex value and corresponds to an M_m-ary modulation symbol, where M_m corresponds to the specific modulation scheme selected for data stream m and $M_m = 2^{q_m}$. Symbol mapping element 316 provides a stream of modulation symbols for each data stream. The modulation symbol streams for all data streams are collectively shown as modulation symbol stream $s(n)$ in FIG. 3.

[1054] Table 1 lists various coding and modulation schemes that may be used to achieve a range of spectral efficiencies (or bit rates) using convolutional and Turbo codes. Each bit rate (in unit of bits/sec/Hertz or bps/Hz) may be achieved using a specific combination of code rate and modulation scheme. For example, a bit rate of one-half may be achieved using a code rate of 1/2 and BPSK modulation, a bit rate of one may be achieved using a code rate of 1/2 and QPSK modulation, and so on.

[1055] In Table 1, BPSK, QPSK, 16-QAM, and 64-QAM are used for the listed bit rates. Other modulation schemes such as DPSK, 8-PSK, 32-QAM, 128-QAM, and so on, may also be used and are within the scope of the invention. DPSK (differential phase-shift keying) may be used when the communication channel is difficult to track since a coherence reference is not needed at the receiver to demodulate a DPSK

modulated signal. For OFDM, modulation may be performed on a per subband basis, and the modulation scheme to be used for each subband may be independently selected.

Table 1

Convolutional Code			**Turbo Code**		
Efficiency (bps/Hz)	Code rate	Modulation	Efficiency (bps/Hz)	Code rate	Modulation
0.5	1/2	BPSK	0.5	1/2	BPSK
1.0	1/2	QPSK	1.0	1/2	QPSK
1.5	3/4	QPSK	1.5	3/4	QPSK
2.0	1/2	16-QAM	2.0	1/2	16-QAM
2.67	2/3	16-QAM	2.5	5/8	16-QAM
3.0	3/4	16-QAM	3.0	3/4	16-QAM
3.5	7/8	16-QAM	3.5	7/12	64-QAM
4.0	2/3	64-QAM	4.0	2/3	64-QAM
4.5	3/4	64-QAM	4.5	3/4	64-QAM
5.0	5/6	64-QAM	5.0	5/6	64-QAM

Other combinations of code rates and modulation schemes may also be used to achieve the various bit rates, and this is also within the scope of the invention.

[1056] In the specific embodiment shown in FIG. 3, transmit processor 220a includes a TX diversity processor 320 and N_T OFDM modulators. Each OFDM modulator includes an inverse fast Fourier transform (IFFT) unit 330 and a cyclic prefix generator 332. TX diversity processor 320 receives and processes the modulation symbols from TX data processor 210a in accordance with one or more selected transmission modes to provide transmit symbols.

[1057] In an embodiment, TX diversity processor 320 further receives and multiplexes pilot symbols (i.e., pilot data) with the transmit symbols using frequency division multiplexing (FDM) in a subset of the available subbands. An example implementation of an FDM pilot transmission scheme is shown in Table 2. In this implementation, 64 subbands are available for the MIMO OFDM system, and subband indices ±7 and ±21 are used for pilot transmission. In alternative embodiments, the pilot symbols may be multiplexed with the transmit symbols using, for example, time

division multiplexing (TDM), code division multiplexing (CDM), or any combination of FDM, TDM, and CDM.

[1058] TX diversity processor 320 provides one transmit symbol stream to each OFDM modulator. The processing by TX diversity processor 320 is described in further detail below.

[1059] Each OFDM modulator receives a respective transmit symbol stream $x_j(n)$. Within each OFDM modulator, IFFT unit 330 groups each set of N_F transmit symbols in stream $x_j(n)$ to form a corresponding symbol vector, and converts the symbol vector into its time-domain representation (which is referred to as an OFDM symbol) using the inverse fast Fourier transform.

[1060] For each OFDM symbol, cyclic prefix generator 332 repeats a portion of the OFDM symbol to form a corresponding transmission symbol. The cyclic prefix ensures that the transmission symbol retains its orthogonal property in the presence of multipath delay spread, thereby improving performance against deleterious path effects such as channel dispersion caused by frequency selective fading. A fixed or an adjustable cyclic prefix may be used for each OFDM symbol. As a specific example of an adjustable cyclic prefix, a system may have a bandwidth of 20 MHz, a chip period of 50 nsec, and 64 subbands. For this system, each OFDM symbol would have a duration of 3.2 μsec (or 64×50 nsec). The cyclic prefix for each OFDM symbol may have a minimum length of 4 chips (200 nsec) and a maximum length of 16 chips (800 nsec), with an increment of 4 chips (200 nsec). Each transmission symbol would then have a duration ranging from 3.4 μsec to 4.0 μsec for cyclic prefixes of 200 nsec to 800 nsec, respectively.

[1061] Cyclic prefix generator 332 in each OFDM modulator provides a stream of transmission symbols to an associated transmitter 222. Each transmitter 222 receives and processes a respective transmission symbol stream to generate a downlink modulated signal, which is then transmitted from the associated antenna 224.

[1062] The coding and modulation for a MIMO OFDM system are described in further detail in the following U.S. patent applications:

- U.S. Patent Application Serial No. 09/993,087, entitled "Multiple-Access Multiple-Input Multiple-Output (MIMO) Communication System," filed November 6, 2001;
- U.S. Patent Application Serial No. 09/854,235, entitled "Method and Apparatus for Processing Data in a Multiple-Input Multiple-Output (MIMO) Communication System Utilizing Channel State Information," filed May 11, 2001;
- U.S. Patent Application Serial Nos. 09/826,481 and 09/956,449, both entitled "Method and Apparatus for Utilizing Channel State Information in a Wireless Communication System," respectively filed March 23, 2001 and September 18, 2001;
- U.S. Patent Application Serial No. 09/776,075, entitled "Coding Scheme for a Wireless Communication System," filed February 1, 2001; and
- U.S. Patent Application Serial No. 09/532,492, entitled "High Efficiency, High Performance Communications System Employing Multi-Carrier Modulation," filed March 30, 2000.

These patent applications are all assigned to the assignee of the present application and incorporated herein by reference.

[1063] The MIMO OFDM system may be designed to support a number of modes of operation for data transmission. These transmission modes include diversity transmission modes, spatial multiplexing transmission modes, and beam steering transmission modes.

[1064] The spatial multiplexing and beam steering modes may be used to achieve higher bit rates under certain favorable channel conditions. These transmission modes are described in further detail in U.S. Patent Application Serial No. 10/085,456, entitled "Multiple-Input, Multiple-Output (MIMO) Systems with Multiple Transmission Modes," filed February 26, 2002, assigned to the assignee of the present application and incorporated herein by reference.

[1065] The diversity transmission modes may be used to achieve higher reliability for certain data transmissions. For example, the diversity transmission modes may be used for overhead channels on the downlink, such as broadcast, paging, and other common channels. The diversity transmission modes may also be used for data

transmission (1) whenever the transmitter does not have adequate channel state information (CSI) for the communication channel, (2) when the channel conditions are sufficiently poor (e.g., under certain mobility conditions) and cannot support more spectrally efficient transmission modes, and (3) for other situations. When the diversity transmission modes are used for downlink data transmission to the terminals, the rate and/or power for each terminal may be controlled to improve performance. A number of diversity transmission modes may be supported and are described in further detail below.

[1066] The diversity transmission modes attempt to achieve transmit diversity by establishing orthogonality among the multiple signals transmitted from multiple transmit antennas. Orthogonality among the transmitted signals may be attained in frequency, time, space, or any combination thereof. Transmit diversity may be established via any one or combination of the following processing techniques:

- Frequency (or subband) diversity. The inherent orthogonality among the subbands provided by OFDM is used to provide diversity against frequency selective fading.
- Transmit diversity using orthogonal functions. Walsh functions or some other orthogonal functions are applied to OFDM symbols transmitted from multiple transmit antennas to establish orthogonality among the transmitted signals. This scheme is also referred to herein as the "Walsh diversity" scheme.
- Space time transmit diversity (STTD). Spatial orthogonality is established between pairs of transmit antennas while preserving the potential for high spectral efficiency offered by MIMO techniques.

[1067] In general, the frequency diversity scheme may be used to combat frequency selective fading and operates in the frequency and spatial dimensions. The Walsh diversity scheme and STTD scheme operate in the time and spatial dimensions.

[1068] For clarity, the processing techniques enumerated above and certain combinations thereof will be described for an example MIMO OFDM system. In this system, each access point is equipped with four antennas to transmit and receive data, and each terminal may be equipped with one or multiple antennas.

Frequency Diversity

[1069] **FIG.** 4 is a block diagram of an embodiment of a TX diversity processor 320a that may be used to implement the frequency diversity scheme. For OFDM, the subbands are inherently orthogonal to one another. Frequency diversity may be established by transmitting identical modulation symbols on multiple subbands.

[1070] As shown in FIG. 4, the modulation symbols, $s(n)$, from TX data processor 210 are provided to a symbol repetition unit 410. Unit 410 repeats each modulation symbol based on the (e.g., dual or quad) diversity to be provided for the modulation symbol. A demultiplexer 412 then receives the repeated symbols and pilot symbols and demultiplexes these symbols into N_T transmit symbol streams. The modulation symbols for each data stream may be transmitted on a respective group of one or more subbands assigned to that data stream. Some of the available subbands may be reserved for pilot transmission (e.g., using FDM). Alternatively, the pilot symbols may be transmitted along with the modulation symbols using TDM or CDM.

[1071] In general, it is desirable to transmit repeated symbols in subbands that are separated from each other by at least the coherence bandwidth of the communication channel. Moreover, the modulation symbols may be repeated over any number of subbands. A higher repetition factor corresponds to greater redundancy and improved likelihood of correct reception at the receiver at the expense of reduced efficiency.

[1072] For clarity, a specific implementation of the frequency diversity scheme is described below for a specific MIMO OFDM system that has some of the characteristics defined by the IEEE Standard 802.11a. The specifications for this IEEE standard are described in a document entitled "Part 11: Wireless LAN Medium Access Control (MAC) and Physical Layer (PHY) specifications: High-speed Physical Layer in the 5 GHz Band," September 1999, which is publicly available and incorporated herein by reference. This system has an OFDM waveform structure with 64 subbands. Of these 64 subbands, 48 subbands (with indices of $\pm\{1, \ldots, 6, 8, \ldots, 20, 22, \ldots, 26\}$) are used for data, 4 subbands (with indices of $\pm\{7, 21\}$) are used for pilot, the DC subband (with index of 0) is not used, and the remaining subbands are also not used and serve as guard subbands.

[1073] Table 2 shows a specific implementation for dual and quad frequency diversity for the system described above. For dual frequency diversity, each modulation

symbol is transmitted over two subbands that are separated by either 26 or 27 subbands. For quad frequency diversity, each modulation symbol is transmitted over four subbands that are separated by 13 or 14 subbands. Other frequency diversity schemes may also be implemented and are within the scope of the invention.

Table 2

Subband Indices	Dual Diversity	Quad Diversity	Subband Indices	Dual Diversity	Quad Diversity
-26	1	1	1	1	1
-25	2	2	2	2	2
-24	3	3	3	3	3
-23	4	4	4	4	4
-22	5	5	5	5	5
-21	pilot	pilot	6	6	6
-20	6	6	7	pilot	pilot
-19	7	7	8	7	7
-18	8	8	9	8	8
-17	9	9	10	9	9
-16	10	10	11	10	10
-15	11	11	12	11	11
-14	12	12	13	12	12
-13	13	1	14	13	1
-12	14	2	15	14	2
-11	15	3	16	15	3
-10	16	4	17	16	4
-9	17	5	18	17	5
-8	18	6	19	18	6
-7	pilot	pilot	20	19	7
-6	19	7	21	pilot	pilot
-5	20	8	22	21	8
-4	21	9	23	22	9
-3	22	10	24	23	10
-2	23	11	25	24	11
-1	24	12	26	25	12
0	DC	DC	-	-	-

[1074] The frequency diversity scheme may be used by a transmitter (e.g., a terminal) not equipped with multiple transmit antennas. In this case, one transmit symbol stream is provided by TX diversity processor 310a. Each modulation symbol in $s(n)$ may be repeated and transmitted on multiple subbands. For single-antenna terminals, frequency diversity may be used to provide robust performance in the presence of frequency selective fading.

[1075] The frequency diversity scheme may also be used when multiple transmit antennas are available. This may be achieved by transmitting the same modulation symbol from all transmit antennas on distinct subbands or subband groups. For example, in a four transmit antenna device, every fourth subband may be assigned to one of the transmit antennas. Each transmit antenna would then be associated with a different group of $N_F/4$ subbands. For quad frequency diversity, each modulation symbol would then be transmitted on a set of four subbands, one in each of the four subband groups, with each group being associated with a specific transmit antenna. The four subbands in the set may also be selected such that they are spaced as far apart as possible. For dual frequency diversity, each modulation may be transmitted on a set of two subbands, one in each of two subband groups. Other implementations for frequency diversity with multiple transmit antennas may also be contemplated, and this is within the scope of the invention. The frequency diversity scheme may also be used in combination with one or more other transmit diversity schemes, as described below.

Walsh Transmit Diversity

[1076] **FIG. 5** is a block diagram of an embodiment of a TX diversity processor 320b that may be used to implement the Walsh diversity scheme. For this diversity scheme, orthogonal functions (or codes) are used to establish time orthogonality, which may in turn be used to establish full transmit diversity across all transmit antennas. This is achieved by repeating the same modulation symbols across the transmit antennas, and time spreading these symbols with a different orthogonal function for each transmit antenna, as described below. In general, various orthogonal functions may be used such as Walsh functions, orthogonal variable spreading factor (OVSF) codes, and so on. For clarity, Walsh functions are used in the following description.

[1077] In the embodiment shown in FIG. 5, the modulation symbols, $s(n)$, from TX data processor 210 are provided to a demultiplexer 510, which demultiplexes the symbols into N_B modulation symbol substreams, one substream for each subband used for data transmission (i.e., each data-carrying subband). Each modulation symbol substream $s_k(n)$ is provided to a respective TX subband processor 520.

[1078] Within each TX subband processor 520, the modulation symbols in substream $s_k(n)$ are provided to N_T multipliers 524a through 524d for the N_T transmit antennas (where $N_T = 4$ for this example system). In the embodiment shown in FIG. 5, one modulation symbol s_k is provided to all four multipliers 524 for each 4-symbol period, which corresponds to a symbol rate of $(4\text{T}_{\text{OFDM}})^{-1}$. Each multiplier also receives a different Walsh function having four chips (i.e., $W_j^4 = \{w_{1j},\ w_{2j},\ w_{3j},\ w_{4j}\}$) and assigned to transmit antenna j associated with that multiplier. Each multiplier then multiplies the symbol s_k with the Walsh function W_j^4 and provides a sequence of four transmit symbols, $\{(s_k \cdot w_{1j}),\ (s_k \cdot w_{2j}),\ (s_k \cdot w_{3j}),\ \text{and}\ (s_k \cdot w_{4j})\}$, which is to be transmitted in four consecutive OFDM symbol periods on subband k of transmit antenna j. These four transmit symbols have the same magnitude as the original modulation symbol s_k. However, the sign of each transmit symbol in the sequence is determined by the sign of the Walsh chip used to generate that transmit symbol. The Walsh function is thus used to time-spread each modulation symbol over four symbol periods. The four multipliers 524a through 524d of each TX subband processor 520 provide four transmit symbol substreams to four buffers/multiplexers 530a through 530d, respectively.

[1079] Each buffer/multiplexer 530 receives pilot symbols and N_B transmit symbol substreams for N_B subbands from N_B TX subband processors 520a through 520f. Each unit 530 then multiplexes the transmit symbols and pilot symbols for each symbol period, and provides a stream of transmit symbols $x_j(n)$ to a corresponding IFFT unit 330. Each IFFT unit 330 receives and processes a respective transmit symbol stream $x_j(n)$ in the manner described above.

[1080] In the embodiment shown in FIG. 5, one modulation symbol is transmitted from all four transmit antennas on each of the N_B data-carrying subbands for each 4-

symbol period. When four transmit antennas are used for data transmission, the spectral efficiency achieved with the Walsh diversity scheme is identical to that achieved with the quad frequency diversity scheme whereby one modulation symbol is transmitted over four data-carrying subbands for each symbol period. In the Walsh diversity scheme with four transmit antennas, the duration or length of the Walsh functions is four OFDM symbols (as designated by the superscript in W_j^4). Since the information in each modulation symbol is distributed over four successive OFDM symbols, the demodulation at the receiver is performed based on four consecutive received OFDM symbols.

[1081] In an alternative embodiment, increased spectral efficiency may be achieved by transmitting distinct modulation symbols (instead of the same modulation symbol) on each transmit antenna. For example, demultiplexer 510 may be designed to provide four distinct modulation symbols, s_1, s_2, s_3, and s_4, to multipliers 524a through 524d for each 4-symbol period. Each multiplier 524 would then multiply a different modulation symbol with its Walsh function to provide a different sequence of four transmit symbols. The spectral efficiency for this embodiment would then be four times that of the embodiment shown in FIG. 5. As another example, demultiplexer 510 may be designed to provide two distinct modulation symbols (e.g., s_1 to multipliers 524a and 524b and s_2 to multipliers 524c and 524d) for each 4-symbol period.

Space-Time Transmit Diversity (STTD)

[1082] Space-time transmit diversity (STTD) supports simultaneous transmission of effectively two independent symbol streams on two transmit antennas while maintaining orthogonality at the receiver. An STTD scheme may thus provide higher spectral efficiency over the Walsh transmit diversity scheme shown in FIG. 5.

[1083] The STTD scheme operates as follows. Suppose that two modulation symbols, denoted as s_1 and s_2, are to be transmitted on a given subband. The transmitter generates two vectors, $\underline{\mathbf{x}}_1 = [s_1 \;\; s_2^*]^T$ and $\underline{\mathbf{x}}_2 = [s_2 \;\; -s_1^*]^T$. Each vector includes two elements that are to be transmitted sequentially in two symbol periods from a respective transmit antenna (i.e., vector $\underline{\mathbf{x}}_1$ is transmitted from antenna 1 and vector $\underline{\mathbf{x}}_2$ is transmitted from antenna 2).

[1084] If the receiver includes a single receive antenna, then the received signal may be expressed in matrix form as:

$$\begin{bmatrix} r_1 \\ r_2 \end{bmatrix} = \begin{bmatrix} h_1 s_1 + h_2 s_2 \\ h_1 s_2^* - h_2 s_1^* \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \end{bmatrix} , \qquad \text{Eq (1)}$$

where r_1 and r_2 are two symbols received in two consecutive symbol periods at the receiver;

h_1 and h_2 are the path gains from the two transmit antennas to the receive antenna for the subband under consideration, where the path gains are assumed to be constant over the subband and static over the 2-symbol period; and

n_1 and n_2 are the noise associated with the two received symbols r_1 and r_2.

[1085] The receiver may then derive estimates of the two transmitted symbols, s_1 and s_2, as follows:

$$\hat{s}_1 = h_1^* r_1 - h_2 r_2^* = (|h_1|^2 + |h_2|^2) s_1 + h_1^* n_1 - h_2 n_2 \text{ , and} \qquad \text{Eq (2)}$$

$$\hat{s}_2 = h_2^* r_1 + h_1 r_2^* = (|h_1|^2 + |h_2|^2) s_2 + h_2^* n_1 + h_1 n_2 \text{ .}$$

[1086] In an alternative implementation, the transmitter may generate two vectors, $\underline{\mathbf{x}}_1 = [s_1 \;\; s_2]^T$ and $\underline{\mathbf{x}}_2 = [-s_2^* \;\; s_1^*]^T$, with the elements of these two vectors being transmitted sequentially in two symbol periods from the two transmit antennas. The received signal may then be expressed as:

$$\begin{bmatrix} r_1 \\ r_2 \end{bmatrix} = \begin{bmatrix} h_1 s_1 - h_2 s_2^* \\ h_1 s_2 + h_2 s_1^* \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \end{bmatrix} .$$

The receiver may then derive estimates of the two transmitted symbols as follows:

$$\hat{s}_1 = h_1^* r_1 + h_2 r_2^* = (|h_1|^2 + |h_2|^2) s_1 + h_1^* n_1 + h_2 n_2 \text{ , and}$$

$$\hat{s}_2 = -h_2 r_1^* + h_1^* r_2 = (|h_1|^2 + |h_2|^2) s_2 - h_2 n_1 + h_1^* n_2 \text{ .}$$

[1087] When two transmit antennas are employed for data transmission, the STTD scheme is twice as spectrally efficient as both the dual frequency diversity scheme and the Walsh diversity scheme with two transmit antennas. The STTD scheme effectively transmits one independent modulation symbol per subband over the two transmit antennas in each symbol period, whereas the dual frequency diversity scheme transmits only a single modulation symbol per two subbands in each symbol period and the Walsh diversity scheme transmits only a single modulation symbol on each subband in two symbol periods. Since the information in each modulation symbol is distributed over two successive OFDM symbols for the STTD scheme, the demodulation at the receiver is performed based on two consecutive received OFDM symbols.

[1088] **FIG. 6** is a block diagram of an embodiment of a TX diversity processor 320c that may be used to implement the STTD scheme. In this embodiment, the modulation symbols, $s(n)$, from TX data processor 210 are provided to a demultiplexer 610, which demultiplexes the symbols into $2N_B$ modulation symbol substreams, two substreams for each data-carrying subband. Each pair of modulation symbol substreams is provided to a respective TX subband processor 620. Each modulation symbol substream includes one modulation symbol for each 2-symbol period, which corresponds to a symbol rate of $(2\text{T}_{\text{OFDM}})^{-1}$.

[1089] Within each TX subband processor 620, the pair of modulation symbol substreams is provided to a space-time encoder 622. For each pair of modulation symbols in the two substreams, space-time encoder 622 provides two vectors, $\underline{\mathbf{x}}_1 = [s_1 \;\; s_2^*]^T$ and $\underline{\mathbf{x}}_2 = [s_2 \;\; -s_1^*]^T$, with each vector including two transmit symbols to be transmitted in two symbol periods. The two transmit symbols in each vector have the same magnitude as the original modulation symbols, s_1 and s_2. However, each transmit symbol may be rotated in phase relative to the original modulation symbol. Each TX subband processor 620 thus provides two transmit symbol substreams to two buffers/multiplexers 630a and 630b, respectively.

[1090] Each buffer/multiplexer 630 receives pilot symbols and N_B transmit symbol substreams from N_B TX subband processors 620a through 620f, multiplexes the transmit symbols and pilot symbols for each symbol period, and provides a stream of transmit

symbols $x_j(n)$ to a corresponding IFFT unit 330. Each IFFT unit 330 then processes a respective transmit symbol stream in the manner described above.

[1091] The STTD scheme is described in further detail by S.M. Alamouti in a paper entitled "A Simple Transmit Diversity Technique for Wireless Communications," IEEE Journal on Selected Areas in Communications, Vol. 16, No. 8, October 1998, pgs. 1451-1458, which is incorporated herein by reference. The STTD scheme is also described in further detail in U.S. Patent Application Serial No. 09/737,602, entitled "Method and System for Increased Bandwidth Efficiency in Multiple Input - Multiple Output Channels," filed January 5, 2001, assigned to the assignee of the present application and incorporated herein by reference.

Walsh-STTD

[1092] A Walsh-STTD scheme employs a combination of Walsh diversity and STTD described above. The Walsh-STTD scheme may be used in systems with more than two transmit antennas. For a Walsh-STTD with repeated symbols scheme (which is also referred to as the repeated Walsh-STTD scheme), two transmit vectors $\underline{\mathbf{x}}_1$ and $\underline{\mathbf{x}}_2$ are generated for each pair of modulation symbols to be transmitted on a given subband from two transmit antennas, as described above for FIG. 6. These two transmit vectors are also repeated across multiple pairs of transmit antennas using Walsh functions to achieve orthogonality across the transmit antenna pairs and to provide additional transmit diversity.

[1093] **FIG. 7** is a block diagram of an embodiment of a TX diversity processor 320d that may be used to implement the repeated Walsh-STTD scheme. The modulation symbols, $s(n)$, from TX data processor 210 are provided to a demultiplexer 710, which demultiplexes the symbols into $2N_B$ modulation symbol substreams, two substreams for each data-carrying subband. Each modulation symbol substream includes one modulation symbol for each 4-symbol period, which corresponds to a symbol rate of $(4T_{OFDM})^{-1}$. Each pair of modulation symbol substreams is provided to a respective TX subband processor 720.

[1094] A space-time encoder 722 within each TX subband processor 720 receives the pair of modulation symbol substreams and, for each 4-symbol period, forms a pair

of modulation symbols $\{s_1$ and $s_2\}$, with one symbol coming from each of the two substreams. The pair of modulation symbols $\{s_1$ and $s_2\}$ is then used to form two vectors, $\underline{\mathbf{x}}_1 = [s_1 \;\; s_2^*]^T$ and $\underline{\mathbf{x}}_2 = [s_2 \;\; -s_1^*]^T$, with each vector spanning a 4-symbol period. Space-time encoder 722 provides the first vector $\underline{\mathbf{x}}_1$ to multipliers 724a and 724c and the second vector $\underline{\mathbf{x}}_2$ to multipliers 724b and 724d. Multipliers 724a and 724b each also receive a Walsh function having two chips (i.e., $W_1^2 = \{w_{11}, \; w_{21}\}$) and assigned to transmit antennas 1 and 2. Similarly, multipliers 724c and 724d each also receive a Walsh function W_2^2 having two chips and assigned to transmit antennas 3 and 4. Each multiplier 724 then multiplies each symbol in its vector $\underline{\mathbf{x}}_j$ with its Walsh function to provide two transmit symbols to be transmitted in two consecutive symbol periods on subband k of transmit antenna j.

[1095] In particular, multiplier 724a multiplies each symbol in vector $\underline{\mathbf{x}}_1$ with the Walsh function W_1^2 and provides a sequence of four transmit symbols, $\{(s_1 \cdot w_{11}),\ (s_1 \cdot w_{21}),\ (s_2^* \cdot w_{11}),$ and $(s_2^* \cdot w_{21})\}$, which is to be transmitted in four consecutive symbol periods. Multiplier 724b multiplies each symbol in vector $\underline{\mathbf{x}}_2$ with the Walsh function W_1^2 and provides a sequence of four transmit symbols, $\{(s_2 \cdot w_{11}),\ (s_2 \cdot w_{21}),\ (-s_1^* \cdot w_{11}),$ and $(-s_1^* \cdot w_{21})\}$. Multiplier 724c multiplies each symbol in vector $\underline{\mathbf{x}}_1$ with the Walsh function W_2^2 and provides a sequence of four transmit symbols, $\{(s_1 \cdot w_{12}),\ (s_1 \cdot w_{22}),\ (s_2^* \cdot w_{12}),$ and $(s_2^* \cdot w_{22})\}$. And multiplier 724d multiplies each symbol in vector $\underline{\mathbf{x}}_2$ with the Walsh function W_2^2 and provides a sequence of four transmit symbols, $\{(s_2 \cdot w_{12}),\ (s_2 \cdot w_{22}),\ (-s_1^* \cdot w_{12}),$ and $(-s_1^* \cdot w_{22})\}$. The Walsh function is thus used to time-spread each symbol or element in the vector $\underline{\mathbf{x}}$ over two symbol periods. The four multipliers 724a through 724d of each TX subband processor 720 provide four transmit symbol substreams to four buffers/multiplexers 730a through 730d, respectively.

[1096] Each buffer/multiplexer 730 receives pilot symbols and N_B transmit symbol substreams from N_B TX subband processors 720a through 720f, multiplexes the pilot and transmit symbols for each symbol period, and provides a stream of transmit

symbols $x_j(n)$ to a corresponding IFFT unit 330. The subsequent processing is as described above.

[1097] The repeated Walsh-STTD scheme shown in FIG. 7 (with four transmit antennas) has the same spectral efficiency as the STTD scheme shown in FIG. 6 and twice the spectral efficiency of the Walsh diversity scheme shown in FIG. 5. However, additional diversity is provided by this Walsh-STTD scheme by transmitting repeated symbols over multiple pairs of transmit antennas. The Walsh-STTD processing provides full transmit diversity (per subband) for the signals transmitted from all transmit antennas.

[1098] **FIG. 8** is a block diagram of an embodiment of a TX diversity processor 320e that may be used to implement a Walsh-STTD without repeated symbols scheme (which is also referred to as the non-repeated Walsh-STTD scheme). This scheme may be used to increase spectral efficiency at the expense of less diversity than the scheme shown in FIG. 7. As shown in FIG. 8, the modulation symbols $s(n)$ are provided to a demultiplexer 810, which demultiplexes the symbols into $4N_B$ modulation symbol substreams, four substreams for each data-carrying subband. Each set of four modulation symbol substreams is provided to a respective TX subband processor 820.

[1099] Within each TX subband processor 820, a space-time encoder 822a receives the first pair of modulation symbol substreams and a space-time encoder 822b receives the second pair of modulation symbol substreams. For each pair of modulation symbols in the two substreams in the first pair, space-time encoder 822a provides two vectors $\underline{\mathbf{x}}_1 = [s_1 \;\; s_2^*]^T$ and $\underline{\mathbf{x}}_2 = [s_2 \;\; -s_1^*]^T$ to multipliers 824a and 824b, respectively. Similarly, for each pair of modulation symbols in the two substreams in the second pair, space-time encoder 822b provides two vectors $\underline{\mathbf{x}}_3 = [s_3 \;\; s_4^*]^T$ and $\underline{\mathbf{x}}_4 = [s_4 \;\; -s_3^*]^T$ to multipliers 824c and 824d, respectively

[1100] Multipliers 824a and 824b each also receive Walsh function W_1^2, and multipliers 824c and 824d each also receive Walsh function W_2^2. Each multiplier 824 then multiplies each symbol in its vector $\underline{\mathbf{x}}_j$ with its Walsh function to provide two transmit symbols to be transmitted in two consecutive symbol periods on subband k of transmit antenna j. The four multipliers 824a through 824d of each TX subband

processor 820 provide four transmit symbol substreams to four buffers/multiplexers 830a through 830d, respectively.

[1101] Each buffer/multiplexer 830 receives pilot symbols and N_B transmit symbol substreams from N_B TX subband processors 820a through 820f, multiplexes the pilot symbols and transmit symbols for each symbol period, and provides a stream of transmit symbols $x_j(n)$ to a corresponding IFFT unit 330. The subsequent processing is as described above.

[1102] The non-repeated Walsh-STTD scheme shown in FIG. 8 (with four transmit antennas) has twice the spectral efficiency as the repeated Walsh-STTD scheme shown in FIG. 7. The same processing may be extended to a system with any number of transmit antenna pairs. Instead of repeating the two transmit vectors across the pairs of transmit antennas, each transmit antenna pair may be used to transmit independent symbol streams. This results in greater spectral efficiency at the possible expense of diversity performance. Some of this diversity may be recovered by the use of forward error correction (FEC) code.

[1103] The Walsh-STTD scheme is also described in further detail in the aforementioned U.S. Patent Application Serial No. 09/737,602.

Frequency-STTD

[1104] A frequency-STTD scheme employs a combination of frequency diversity and STTD. The frequency-STTD scheme may also employ antenna diversity for systems with more than one pair of transmit antennas. For the frequency-STTD scheme, each modulation symbol is transmitted on multiple (e.g., two) subbands and provided to multiple TX subband processors. The subbands to be used for each modulation symbol may be selected such that they are spaced as far apart as possible (e.g., as shown in Table 1) or based on some other subband assignment scheme. If four transmit antennas are available, then for each subband two pairs of modulation symbols are processed using STTD. The first pair of modulation symbols is transmitted from the first pair of antennas (e.g., transmit antennas 1 and 2), and the second pair of modulation symbols is transmitted from the second pair of antennas (e.g., transmit antennas 3 and 4).

[1105] Each modulation symbol is thus transmitted on multiple subbands and over multiple transmit antennas. For clarity, the processing for a given modulation symbol s_a for a system with four transmit antennas and using dual frequency diversity may be performed as follows. Modulation symbol s_a is initially provided to two TX subband processors (e.g., for subbands k and $k+N_F/2$). In subband k, modulation symbol s_a is processed with another modulation symbol s_b using STTD to form two vectors, $\underline{\mathbf{x}}_1 = [s_a \ \ s_b^*]^T$ and $\underline{\mathbf{x}}_2 = [s_b \ \ -s_a^*]^T$, which are transmitted from transmit antennas 1 and 2, respectively. In subband $k+N_F/2$, modulation symbol s_a is processed with another modulation symbol s_c using STTD to form two vectors, $\underline{\mathbf{x}}_3 = [s_a \ \ s_c^*]^T$ and $\underline{\mathbf{x}}_4 = [s_c \ \ -s_a^*]^T$, which are transmitted from transmit antennas 3 and 4, respectively. Modulation symbol s_c may be the same as modulation symbol s_b or a different modulation symbol.

[1106] For the above implementation of the frequency-STTD scheme, the modulation symbol in each subband has two orders of transmit diversity provided by the STTD processing. Each modulation symbol to be transmitted has four orders of transmit diversity plus some frequency diversity provided by the use of two subbands and STTD. This frequency-STTD scheme has the same spectral efficiency as the repeated Walsh-STTD scheme. However, the total transmission time for each modulation symbol is two symbol periods with the frequency-STTD scheme, which is half the total transmission time for each modulation symbol with the Walsh-STTD scheme, since Walsh processing is not performed by the frequency-STTD scheme.

[1107] In one embodiment of the frequency-STTD scheme, all subbands are used by each pair of transmit antennas for data transmission. For quad diversity, each modulation symbol is provided to two subbands for two transmit antenna pairs, as described above. In another embodiment of the frequency-STTD scheme, each pair of transmit antennas is assigned a different subband group for data transmission. For example, in a device with two pairs of transmit antennas, every other subband may be assigned to one transmit antenna pair. Each transmit antenna pair would then be associated with a different group of $N_F/2$ subbands. For quad diversity, each modulation symbol would then be transmitted on two subbands, one in each of the two

subband groups, with each group being associated with a specific transmit antenna pair. The two subbands used for each modulation symbol may be selected such that they are spaced as far apart as possible. Other implementations for frequency-STTD diversity with multiple transmit antenna pairs may also be contemplated, and this is within the scope of the invention.

[1108] As illustrated by the above, various diversity schemes may be implemented using various processing techniques described herein. For clarity, specific implementations of various diversity schemes have been described above for a specific system. Variations of these diversity schemes may also be implemented, and this is within the scope of the invention.

[1109] Moreover, other diversity schemes may also be implemented based on other combinations of the processing techniques described herein, and this is also within the scope of the invention. For example, another diversity scheme may utilize frequency diversity and Walsh transmit diversity, and yet another diversity scheme may utilize frequency diversity, Walsh diversity, and STTD.

Diversity Transmission Modes

[1110] A number of diversity transmission modes may be implemented using the transmit processing schemes described above. These diversity transmission modes may include the following:

- Frequency diversity transmission mode - employs only frequency diversity (e.g., dual, quad, or some other integer multiple frequency diversity).
- Walsh diversity transmission mode - employs only Walsh transmit diversity.
- STTD transmission mode - employs only STTD.
- Walsh-STTD transmission mode - employs both Walsh transmit diversity and STTD, with repeated or non-repeated symbols.
- Frequency-STTD transmission mode - employs frequency diversity and STTD.
- Frequency-Walsh transmission mode - employs frequency diversity and Walsh transmit diversity.

- Frequency-Walsh-STTD transmission mode - employs frequency diversity. Walsh transmit diversity, and STTD.

[1111] The diversity transmission modes may be used for data transmission between the access points and terminals. The specific transmission mode to use for a given data stream may be dependent on various factors such as (1) the type of data being transmitted (e.g., whether common for all terminals or user-specific for a particular terminal), (2) the number of antennas available at the transmitter and receiver, (3) the channel conditions, (4) the requirements of the data transmission (e.g., the required packet error rate), and so on.

[1112] Each access point in the system may be equipped with, for example, four antennas for data transmission and reception. Each terminal may be equipped with one, two, four, or some other number of antennas for data transmission and reception. Default diversity transmission modes may be defined and used for each terminal type. In a specific embodiment, the following diversity transmission modes are used as default:

- Single-antenna terminals – use frequency diversity transmission mode with dual or quad diversity.
- Dual-antenna terminals – use STTD transmission mode for dual diversity and frequency-STTD transmission mode for quad diversity.
- Quad-antenna terminals – use STTD transmission mode for dual diversity and Walsh-STTD transmission mode for quad diversity.

Other diversity transmission modes may also be selected as the default modes, and this is within the scope of the invention.

[1113] The diversity transmission modes may also be used to increase the reliability of data transmission on overhead channels intended to be received by all terminals in the system. In an embodiment, a specific diversity transmission mode is used for the broadcast channel, and this mode is known *a priori* by all terminals in the system (i.e., no signaling is required to identify the transmission mode used for the broadcast channel). In this way, the terminals are able to process and recover the data transmitted on the broadcast channel. The transmission modes used for other overhead channels may be fixed or dynamically selected. In one dynamic selection scheme, the system

determines which transmission mode is the most reliable (and spectrally efficient) to use for each of the remaining overhead channels based on the mix of terminals being served. The transmission modes selected for use for these overhead channels and other configuration information may be signaled to the terminals, for example, via the broadcast channel.

[1114] With OFDM, the subbands may be treated as distinct transmission channels, and the same or different diversity transmission modes may be used for the subbands. For example, one diversity transmission mode may be used for all data-carrying subbands, or a separate diversity transmission mode may be selected for each data-carrying subband. Moreover, for a given subband, it may be possible to use different diversity transmission modes for different sets of transmit antennas.

[1115] In general, each data stream (whether for an overhead channel or a specific receiver device) may be coded and modulated based on the coding and modulation schemes selected for that data stream to provide modulation symbols. The modulation symbols are then further processed based on the diversity transmission mode selected for that data stream to provide transmit symbols. The transmit symbols are further processed and transmitted on a group of one or more subbands from a set of one or more transmit antennas designated to be used for that data stream.

Receiver Unit

[1116] **FIG. 9** is a block diagram of a receiver unit 900, which is an embodiment of the receiver portion of a multi-antenna terminal 106. The downlink modulated signals from access point 104 are received by antennas 252a through 252r, and the received signal from each antenna is provided to a respective receiver 254. Each receiver 254 processes (e.g., conditions, digitizes, and data demodulates) the received signal to provide a stream of received transmission symbols, which is then provided to a respective OFDM demodulator within a receive processor 260a.

[1117] Each OFDM demodulator includes a cyclic prefix removal unit 912 and a fast Fourier transform (FFT) unit 914. Unit 912 removes the cyclic prefix that had been appended in each transmission symbol to provide a corresponding received OFDM symbol. The cyclic prefix removal may be performed by determining a set of N_A samples corresponding to each received transmission symbol and selecting a subset of

these N_A samples as the set of N_F samples for the received OFDM symbol. FFT 914 then transforms each received OFDM symbol (or each set of N_F samples) using the fast Fourier transform to provide a vector of N_F received symbols for the N_F subbands. FFT units 914a through 914r provide N_R received symbol streams, $r_1(n)$ through $r_{N_R}(n)$, to an RX diversity processor 920.

[1118] RX diversity processor 920 performs diversity processing on the N_R received symbol streams to provide recovered symbols, $\hat{s}(n)$, which are estimates of the modulation symbols, $s(n)$, sent by the transmitter. The processing to be performed by RX diversity processor 920 is dependent on the transmission mode used for each data stream to be recovered, as indicated by the transmission mode control. RX diversity processor 920 is described in further detail below.

[1119] RX diversity processor 920 provides the recovered symbols, $\hat{s}(n)$, for all data streams to be recovered to an RX data processor 262a, which is an embodiment of RX data processor 262 in FIG. 2. Within processor 262a, a symbol demapping element 942 demodulates the recovered symbols for each data stream in accordance with a demodulation scheme that is complementary to the modulation scheme used for the data stream. A channel deinterleaver 944 then deinterleaves the demodulated data in a manner complementary to the interleaving performed at the transmitter for the data stream, and the deinterleaved data is further decoded by a decoder 946 in a manner complementary to the coding performed at the transmitter. For example, a Turbo decoder or a Viterbi decoder may be used for decoder 946 if Turbo or convolutional coding, respectively, is performed at the transmitter. The decoded data from decoder 946 represents an estimate of the transmitted data being recovered. Decoder 946 may also provide the status of each received packet (e.g., indicating whether it was received correctly or in error).

[1120] In the embodiment shown in FIG. 9, a channel estimator 950 estimates various channel characteristics such as the channel response and the noise variance (e.g., based on recovered pilot symbols) and provides these estimates to controller 270. Controller 270 may be designed to perform various functions related to diversity processing at the receiver. For example, controller 270 may determine the diversity transmission mode used for each data stream to be recovered and may further direct the operation of RX diversity processor 920.

[1121] **FIG. 10** is a block diagram of an embodiment of an RX diversity processor 920x, which may be used for a multi-antenna receiver device. In this embodiment, the N_R received symbol streams for the N_R receive antennas are provided to N_R RX antenna processors 1020a through 1020r. Each RX antenna processor 1020 processes a respective received symbol stream, $r_i(n)$, and provides a corresponding recovered symbol stream, $\hat{s}_i(n)$, for the associated receive antenna. In an alternative embodiment, one or more RX antenna processors 1020 are time shared and used to process all N_R received symbol streams.

[1122] A combiner 1030 then receives and combines the N_R recovered symbol streams from the N_R RX antenna processors 1020a through 1020r to provide a single recovered symbol stream, $\hat{s}(n)$. The combining may be performed on a symbol-by-symbol basis. In an embodiment, for a given subband k, the N_R recovered symbols from the N_R receive antennas for each symbol period (which are denoted as $\{\hat{s}_{ki}\}$, for $i = (1, 2, \ldots, N_R)$) are initially scaled by N_R weights assigned to the N_R receive antennas. The N_R scaled symbols are then summed to provide the recovered symbol, $\hat{s}_k$, for subband k. The weights may be selected to achieve maximal-ratio combining, and may be determined based on the signal quality (e.g., SNR) associated with the receive antennas. The scaling with the weights may also be performed via an automatic gain control (AGC) loop maintained for each receive antenna, as is known in the art.

[1123] For a single-antenna receiver device, there is only one received symbol stream. In this case, only one RX antenna processor 1020 is needed. A design for RX antenna processor 1020 is described in further detail below.

[1124] The recovered symbol steam, $\hat{s}(n)$, provided by combiner 1030 may include the recovered symbols for all data streams transmitted by the transmitter. Alternatively, the steam $\hat{s}(n)$ may include only the recovered symbols for one or more data streams to be recovered by the receiver device.

[1125] **FIG. 11** is a block diagram of an RX antenna processor 1020x that may be used to perform the receive processing for the Walsh diversity scheme shown in FIG. 5. RX antenna processor 1020x processes the received symbol stream $r_i(n)$ for one receive antenna and may be used for each of RX antenna processors 1020a through 1020r in FIG. 10.

[1126] In the embodiment shown in FIG. 11, the received symbol stream $r_i(n)$ is provided to a demultiplexer 1110, which demultiplexes the received symbols in $r_i(n)$ into N_B substreams of received symbols (which are denoted as r_1 through r_{N_B}, where the index i has been dropped for simplicity), one substream for each data-carrying subband. Each received symbol substream r_k is then provided to a respective RX subband processor 1120.

[1127] Each RX subband processor 1120 includes a number of receive processing paths, one path for each transmit antenna used for data transmission (four receive processing paths are shown in FIG. 11 for four transmit antennas). For each processing path, the received symbols in the substream are provided to a multiplier 1122 that also receives a scaled Walsh function $\hat{h}_{kj}^{*}(W_j^4)^*$, where $\hat{h}_{kj}^{*}$ is the complex-conjugated channel response estimate between transmit antenna j (which is associated with that multiplier) and the receive antenna for subband k, and $(W_j^4)^*$ is the complex-conjugated Walsh function assigned to transmit antenna j. Each multiplier 1122 then multiplies the received symbols with the scaled Walsh function and provides the results to an associated integrator 1124. Integrator 1124 then integrates the multiplier results over the length of the Walsh function (or four symbol periods) and provides the integrated output to a summer 1126. One received symbol is provided to multiplier 1122 for each symbol period (i.e., rate $= (T_{OFDM})^{-1}$) and integrator 1124 provides one integrated output for each 4-symbol period (i.e., rate $= (4T_{OFDM})^{-1}$).

[1128] For each 4-symbol period, summer 1126 combines the four outputs from integrators 1124a through 1124d to provide a recovered symbol, $\hat{s}_k$, for subband k, which is an estimate of the modulation symbol, s_k, transmitted in that subband. For each 4-symbol period, RX subband processors 1120a through 1120f provide N_B recovered symbols, $\hat{s}_1$ through $\hat{s}_{N_B}$, for the N_B data-carrying subbands.

[1129] A multiplexer 1140 receives the recovered symbols from RX subband processors 1120a through 1120f and multiplexes these symbols into a recovered symbol stream, $\hat{s}_i(n)$, for receive antenna i.

[1130] **FIG. 12** is a block diagram of an RX subband processor 1120x that may be used to perform the receive processing for the Walsh-STTD schemes shown in FIGS. 7

and 8. RX subband processor 1120x processes one received symbol substream r_k for one subband of one receive antenna and may be used for each of RX subband processors 1120a through 1120f in FIG. 11.

[1131] In the embodiment shown in FIG. 12, the received symbols in substream r_k are provided to two receive processing paths, one path for each transmit antenna pair used for data transmission (two receive processing paths are shown in FIG. 12 for four transmit antennas). For each processing path, the received symbols are provided to a multiplier 1222 that also receives a complex-conjugated Walsh function $(W_j^2)^*$ assigned to the transmit antenna pair being processed by that path. Each multiplier 1222 then multiplies the received symbols with the Walsh function and provides the results to an associated integrator 1224. Integrator 1224 then integrates the multiplier results over the length of the Walsh function (or two symbol periods) and provides the integrated output to a delay element 1226 and a unit 1228. One received symbol is provided to multiplier 1222 for each symbol period (i.e., rate $= (\mathrm{T_{OFDM}})^{-1}$) and integrator 1224 provides one integrated output for each 2-symbol period (i.e., rate $= (2\mathrm{T_{OFDM}})^{-1}$).

[1132] Referring back to FIG. 8, for the non-repeated Walsh-STTD scheme, four modulation symbols $\{s_{k1}$, s_{k2}, s_{k3}, and $s_{k4}\}$ are transmitted over two transmit antenna pairs in four symbol periods for subband k (where the index k is used to denote subband k). The symbol pair $\{s_{k1}$ and $s_{k2}\}$ is transmitted over the first transmit antenna pair, and the symbol pair $\{s_{k3}$ and $s_{k4}\}$ is transmitted over the second transmit antenna pair. Each modulation symbol is transmitted in two symbol periods using the 2-chip Walsh function assigned to the transmit antenna pair.

[1133] Referring back to FIG. 12, the complementary processing is performed at the receiver to recover the modulation symbols. For each 4-symbol period corresponding to a new symbol pair transmitted from each transmit antenna pair for subband k, integrator 1224 provides a received symbol pair $\{r_{k1}$ and $r_{k2}\}$. Delay element 1226 then provides a delay of two symbol periods (i.e., $\mathrm{T_W} = 2\mathrm{T_{OFDM}}$, which is the length of the Walsh function) for the first symbol (i.e., r_{k1}) in the pair, and unit 1228 provides the complex-conjugate of the second symbol (i.e., r_{k2}^*) in the pair.

[1134] Multipliers 1230a through 1230d and summers 1232a and 1232b then collectively perform the computations shown in equation (2) for the first transmit antenna pair. In particular, multiplier 1230a multiplies the symbol r_{k1} with the channel response estimate $\hat{h}^*_{k1}$, multiplier 1230b multiplies the symbol r^*_{k2} with the channel response estimate $\hat{h}_{k2}$, multiplier 1230c multiplies the symbol r_{k1} with the channel response estimate $\hat{h}^*_{k2}$, and multiplier 1230d multiplies the symbol r^*_{k2} with the channel response estimate $\hat{h}_{k1}$, where $\hat{h}_{kj}$ is an estimate of the channel response from transmit antenna j to the receive antenna for subband k. Summer 1232a then subtracts the output of multiplier 1230b from the output of multiplier 1230a to provide an estimate, $\hat{s}_{k1}$, of the first modulation symbol in the pair $\{s_{k1}$ and $s_{k2}\}$. Summer 1232b adds the output of multiplier 1230c with the output of multiplier 1230d to provide an estimate, $\hat{s}_{k2}$, of the second modulation symbol in the pair.

[1135] The processing by the second path for the second transmit antenna pair is similar to that described above for the first path. However, the channel response estimates, $\hat{h}_{k3}$ and $\hat{h}_{k4}$, for the second pair of transmit antennas for subband k are used for the second processing path. For each 4-symbol period, the second processing path provides the symbol estimates $\hat{s}_{k3}$ and $\hat{s}_{k4}$ for the pair of modulation symbols $\{s_{k3}$ and $s_{k4}\}$ transmitted on subband k from the second transmit antenna pair.

[1136] For the non-repeated Walsh-STTD scheme shown in FIG. 8, $\hat{s}_{k1}$, $\hat{s}_{k2}$, $\hat{s}_{k3}$, and $\hat{s}_{k4}$ represent the estimates of the four modulation symbols s_{k1}, s_{k2}, s_{k3}, and s_{k4} sent over four transmit antennas on subband k in a 4-symbol period. These symbol estimates may then be multiplexed together into a recovered symbol substream, $\hat{s}_k(n)$, for subband k, which is then provided to multiplexer 1140 in FIG. 11.

[1137] For the repeated Walsh-STTD scheme shown in FIG. 7, one symbol pair $\{s_{k1}$ and $s_{k2}\}$ is sent over both pairs of transmit antennas on subband k in each 4-symbol period. The symbol estimates $\hat{s}_{k1}$ and $\hat{s}_{k3}$ may then be combined by a summer (not shown in FIG. 12) to provide an estimate of the first symbol in the pair, and the

symbol estimates $\hat{s}_{k2}$ and $\hat{s}_{k4}$ may similarly be combined by another summer to provide an estimate of the second symbol in the pair. The symbol estimates from these two summers may then be multiplexed together into a recovered symbol substream, $\hat{s}_k(n)$, for subband k, which is then provided to multiplexer 1140 in FIG. 11.

[1138] For clarity, various details are specifically described for downlink data transmission from an access point to a terminal. The techniques described herein may also be used for the uplink, and this is within the scope of the invention. For example, the processing schemes shown in FIGS. 4, 5, 6, 7, and 8 may be implemented within a multi-antenna terminal for uplink data transmission.

[1139] The MIMO OFDM system described herein may also be designed to implement one or more multiple access schemes such as code division multiple access (CDMA), time division multiple access (TDMA), frequency division multiple access (FDMA), and so on. CDMA may provide certain advantages over other types of system, such as increased system capacity. The MIMO OFDM system may also be designed to implement various processing techniques described in CDMA standards such as IS-95, cdma2000, IS-856, W-CDMA, and others.

[1140] The techniques described herein for transmitting and receiving data using a number of diversity transmission modes may be implemented by various means. For example, these techniques may be implemented in hardware, software, or a combination thereof. For a hardware implementation, the elements (e.g., TX diversity processor, RX diversity processor, TX subband processors, RX antenna processors, RX subband processors, and so on) used to implement any one or a combination of the techniques may be implemented within one or more application specific integrated circuits (ASICs), digital signal processors (DSPs), digital signal processing devices (DSPDs), programmable logic devices (PLDs), field programmable gate arrays (FPGAs), processors, controllers, micro-controllers, microprocessors, other electronic units designed to perform the functions described herein, or a combination thereof.

[1141] For a software implementation, any one or a combination of the techniques described herein may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memory 232 or 272 in FIG. 2) and executed by a processor (e.g., controller 230 or 270). The memory unit may be implemented within the processor or

external to the processor, in which case it can be communicatively coupled to the processor via various means as it known in the art.

[1142] Headings are included herein for reference and to aid in locating certain sections. These heading are not intended to limit the scope of the concepts described therein under, and these concepts may have applicability in other sections throughout the entire specification.

[1143] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[1144] WHAT IS CLAIMED IS:

CLAIMS

1. A method for processing data for transmission in a wireless communication system, comprising:

selecting a particular diversity transmission mode from among a plurality of possible transmission modes to use for each of one or more data streams, wherein each selected diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof;

coding and modulating each data stream based on coding and modulation schemes selected for the data stream to provide modulation symbols; and

processing the modulation symbols for each data stream based on the selected diversity transmission mode to provide transmit symbols for transmission over one or more transmit antennas.

2. The method of claim 1, wherein the plurality of possible transmission modes includes a frequency diversity transmission mode.

3. The method of claim 1, wherein the plurality of possible transmission modes includes a Walsh diversity transmission mode.

4. The method of claim 3, wherein the Walsh diversity transmission mode transmits each modulation symbol over N_T symbol periods, where N_T is the number of transmit antennas used for data transmission.

5. The method of claim 4, wherein the Walsh diversity transmission mode transmits each modulation symbol over all N_T transmit antennas.

6. The method of claim 1, wherein the plurality of possible transmission modes includes a space time transmit diversity (STTD) transmission mode.

7. The method of claim 1, wherein the plurality of possible transmission modes includes a Walsh-STTD transmission mode.

8. The method of claim 1, wherein the plurality of possible transmission modes includes a frequency-STTD transmission mode.

9. The method of claim 7, wherein the Walsh-STTD transmission mode redundantly transmits modulation symbols over a plurality of pairs of transmit antennas.

10. The method of claim 7, wherein the Walsh-STTD transmission mode transmits different modulation symbols over different pairs of transmit antennas.

11. The method of claim 1, wherein the wireless communication system is a multiple-input multiple-output (MIMO) communication system, and wherein the transmit symbols for the one or more data streams are transmitted over a plurality of transmit antennas.

12. The method of claim 11, wherein the MIMO communication system utilizes orthogonal frequency division multiplexing (OFDM).

13. The method of claim 12, further comprising:

OFDM modulating the transmit symbols for the one or more data streams to provide a stream of transmission symbols for each transmit antenna used for data transmission.

14. The method of claim 12, wherein the transmit symbols for each data stream are transmitted on a respective group of one or more subbands.

15. The method of claim 1, wherein at least one data stream is transmitted for an overhead channel.

16. The method of claim 14, wherein the data stream for a broadcast channel is transmitted based on a fixed diversity transmission mode.

17. The method of claim 1, wherein at least one data stream is user-specific and transmitted for a specific receiver device.

18. The method of claim 17, wherein data rate for each of the at least one user-specific data stream is adjusted based on transmission capability of the specific receiver device.

19. The method of claim 1, further comprising:

multiplexing pilot symbols with the modulation symbols for the one or more data streams.

20. The method of claim 1, wherein the pilot symbols are multiplexed with the modulation symbols using frequency division multiplexing (FDM).

21. A method for processing data for transmission in a multiple-input multiple-output (MIMO) communication system that utilizes orthogonal frequency division multiplexing (OFDM), comprising:

selecting a particular diversity transmission mode from among a plurality of possible transmission modes to use for each of one or more data streams, wherein each selected diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof by using frequency diversity, Walsh transmit diversity, space time transmit diversity (STTD), or any combination thereof;

coding and modulating each data stream based on coding and modulation schemes selected for the data stream to provide modulation symbols; and

processing the modulation symbols for each data stream based on the selected diversity transmission mode to provide transmit symbols for transmission over a plurality of transmit antennas.

22. The method of claim 21, wherein the plurality of possible transmission modes includes a frequency diversity transmission mode, a Walsh diversity transmission mode, and a STTD transmission mode.

23. The method of claim 22, wherein the plurality of possible transmission modes further includes a Walsh-STTD transmission mode.

24. A method for processing data for transmission in a multiple-input multiple-output (MIMO) communication system that utilizes orthogonal frequency division multiplexing (OFDM), comprising:

coding and modulating data to provide one or more substreams of modulation symbols for each of a plurality of OFDM subbands; and

for each of the plurality of OFDM subbands, processing the modulation symbols in the one or more substreams for the OFDM subband to provide transmit symbols, wherein the modulation symbols are processed in accordance with a particular diversity processing scheme selected for the OFDM subband to provide diversity in time, frequency, space, or a combination thereof.

25. The method of claim 24, wherein the selected diversity processing scheme for at least one OFDM subband is a space time transmit diversity (STTD) scheme.

26. The method of claim 24, wherein the selected diversity processing scheme for at least one OFDM subband is a Walsh transmit diversity scheme.

27. The method of claim 24, wherein the selected diversity processing scheme for at least one OFDM subband is a Walsh-space time transmit diversity (Walsh-STTD) scheme.

28. A method for processing a data transmission at a receiver of a wireless communication system, comprising:

determining a particular diversity transmission mode used for each of one or more data streams to be recovered from a received data transmission, wherein the diversity transmission mode used for each data stream is selected from among a plurality of possible transmission modes, and wherein each diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof; and

processing received symbols for each data stream based on the diversity transmission mode used for the data stream to provide recovered symbols that are estimates of modulation symbols transmitted from a transmitter for the data stream.

29. The method of claim 28, wherein the plurality of possible transmission modes includes a frequency diversity transmission mode, a Walsh diversity transmission mode, and a space time transmit diversity (STTD) transmission mode.

30. The method of claim 29, wherein the plurality of possible transmission modes further includes a Walsh-STTD transmission mode.

31. The method of claim 29, wherein the plurality of possible transmission modes further includes a frequency-STTD transmission mode.

32. The method of claim 28, further comprising:

demodulating and decoding the recovered symbols for each data stream to provide decoded data.

33. A memory communicatively coupled to a digital signal processing device (DSPD) capable of interpreting digital information to:

select a particular diversity transmission mode from among a plurality of possible transmission modes to use for each of one or more data streams, wherein each selected diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof;

code and modulate each data stream based on coding and modulation schemes selected for the data stream to provide modulation symbols; and

process the modulation symbols for each data stream based on the selected diversity transmission mode to provide transmit symbols for transmission over one or more transmit antennas.

34. A transmitter unit in a wireless communication system, comprising:

a controller operative to select a particular diversity transmission mode from among a plurality of possible transmission modes to use for each of one or more data streams, wherein each selected diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof;

a TX data processor operative to code and modulate each data stream based on coding and modulation schemes selected for the data stream to provide modulation symbols; and

a transmit processor operative to process the modulation symbols for each data stream based on the selected diversity transmission mode to provide transmit symbols for transmission over one or more transmit antennas.

35. The transmitter unit of claim 34, wherein the plurality of possible transmission modes includes a frequency diversity transmission mode, a Walsh diversity transmission mode, and a space time transmit diversity (STTD) transmission mode.

36. The transmitter unit of claim 35, wherein the plurality of possible transmission modes further includes a Walsh-STTD transmission mode.

37. The transmitter unit of claim 35, wherein the plurality of possible transmission modes further includes a frequency-STTD transmission mode.

38. The transmitter unit of claim 34, wherein the wireless communication system is a multiple-input multiple-output (MIMO) communication system that utilizes orthogonal frequency division multiplexing (OFDM).

39. The transmitter unit of claim 38, wherein the transmit processor is further operative to OFDM modulate the transmit symbols for the one or more data streams to provide a stream of transmission symbols for each transmit antenna used for data transmission.

40. An access point comprising the transmitter unit of claim 34.

41. A terminal comprising the transmitter unit of claim 34.

42. An apparatus in a multiple-input multiple-output (MIMO) communication system, comprising:

means for selecting a particular diversity transmission mode from among a plurality of possible transmission modes to use for each of one or more data streams, wherein each selected diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof;

means for coding and modulating each data stream based on coding and modulation schemes selected for the data stream to provide modulation symbols; and

means for processing the modulation symbols for each data stream based on the selected diversity transmission mode to provide transmit symbols for transmission over one or more transmit antennas.

43. The apparatus of claim 42, wherein the plurality of possible transmission modes includes a frequency diversity transmission mode, a Walsh diversity transmission mode, and a space time transmit diversity (STTD) transmission mode.

44. A receiver unit in a wireless communication system, comprising:

a controller operative to determine a particular diversity transmission mode used for each of one or more data streams to be recovered from a received data transmission, wherein the diversity transmission mode used for each data stream is selected from among a plurality of possible transmission modes, and wherein each diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof; and

a receive processor operative to process received symbols for each data stream based on the diversity transmission mode used for the data stream to provide recovered symbols that are estimates of modulation symbols transmitted from a transmitter for the data stream.

45. The receiver unit of claim 44, further comprising:

a receive data processor operative to demodulate and decode the recovered symbols for each data stream to provide decoded data.

46. The receiver unit of claim 44, wherein the plurality of possible transmission modes includes a frequency diversity transmission mode, a Walsh

diversity transmission mode, and a space time transmit diversity (STTD) transmission mode.

47. An access point comprising the receiver unit of claim 44.

48. A terminal comprising the receiver unit of claim 44.

49. A receiver apparatus in a wireless communication system, comprising:

means for determining a particular diversity transmission mode used for each of one or more data streams to be recovered from a received data transmission, wherein the diversity transmission mode used for each data stream is selected from among a plurality of possible transmission modes, and wherein each diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof; and

means for processing received symbols for each data stream based on the diversity transmission mode used for the data stream to provide recovered symbols that are estimates of modulation symbols transmitted from a transmitter for the data stream.

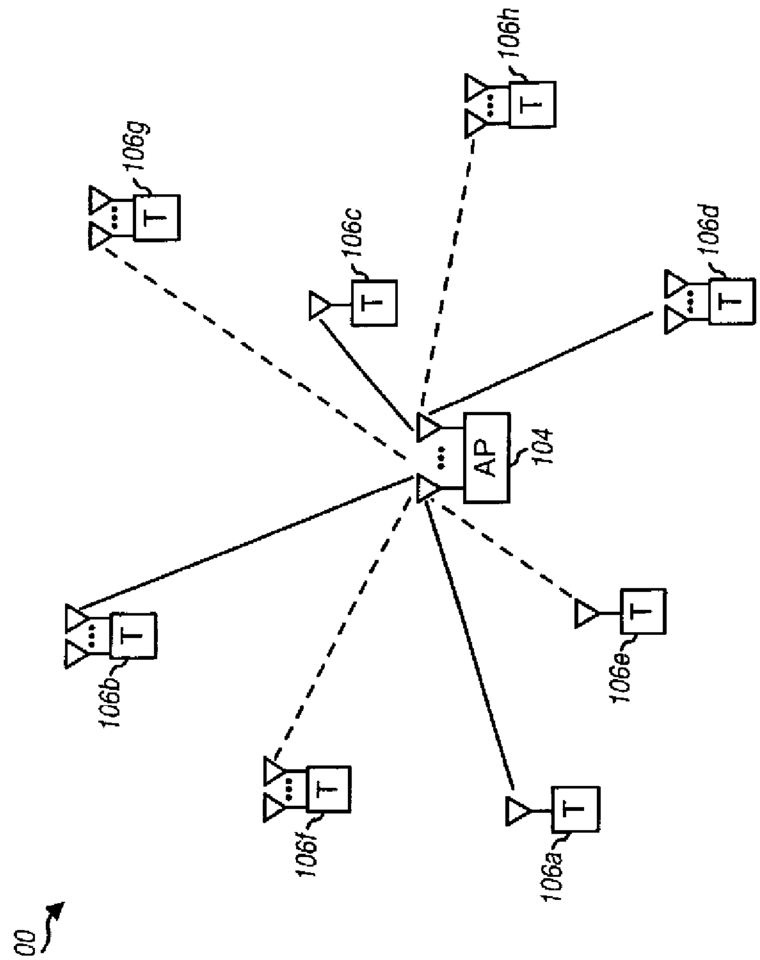


FIG. 1

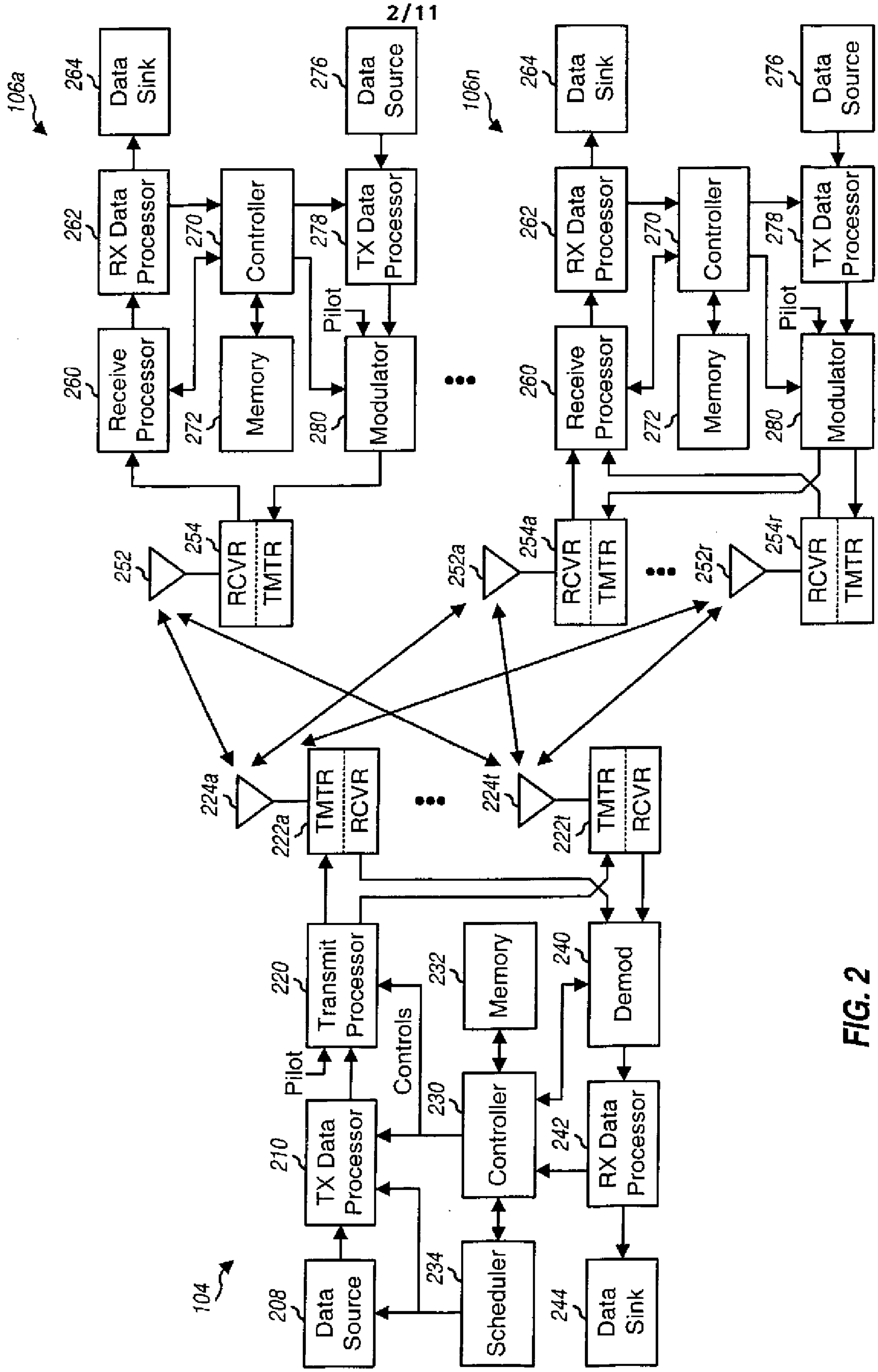

106a
264
Data Sink
262
RX Data Processor
270
Controller
276
Data Source
278
TX Data Processor
260
Receive Processor
272
Memory
Pilot
280
Modulator
252
254
RCVR
TMTR
106n
264
Data Sink
262
RX Data Processor
270
Controller
276
Data Source
278
TX Data Processor
260
Receive Processor
272
Memory
Pilot
280
Modulator
252a
254a
RCVR
TMTR
252r
254r
RCVR
TMTR
224a
222a
TMTR
RCVR
224t
222t
TMTR
RCVR
104
208
Data Source
210
TX Data Processor
Pilot
220
Transmit Processor
Controls
232
Memory
230
Controller
234
Scheduler
240
Demod
242
RX Data Processor
244
Data Sink

FIG. 2

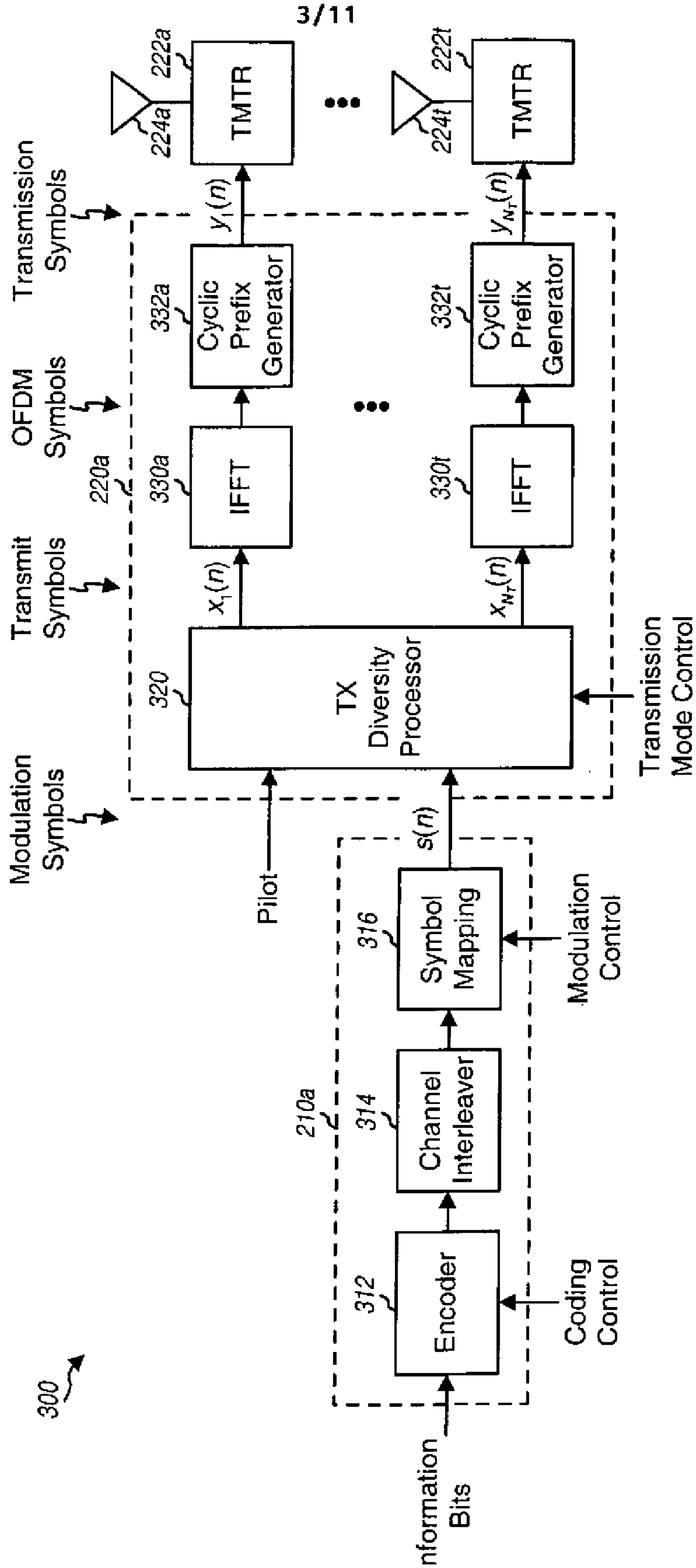

300
Information Bits
210a
312
Encoder
Coding Control
314
Channel Interleaver
316
Symbol Mapping
Modulation Control
s(n)
Pilot
Modulation Symbols
220a
320
TX Diversity Processor
Transmission Mode Control
Transmit Symbols
x1(n)
xNT(n)
330a
IFFT
330t
IFFT
OFDM Symbols
332a
Cyclic Prefix Generator
332t
Cyclic Prefix Generator
Transmission Symbols
y1(n)
yNT(n)
222a
TMTR
224a
222t
TMTR
224t

FIG. 3

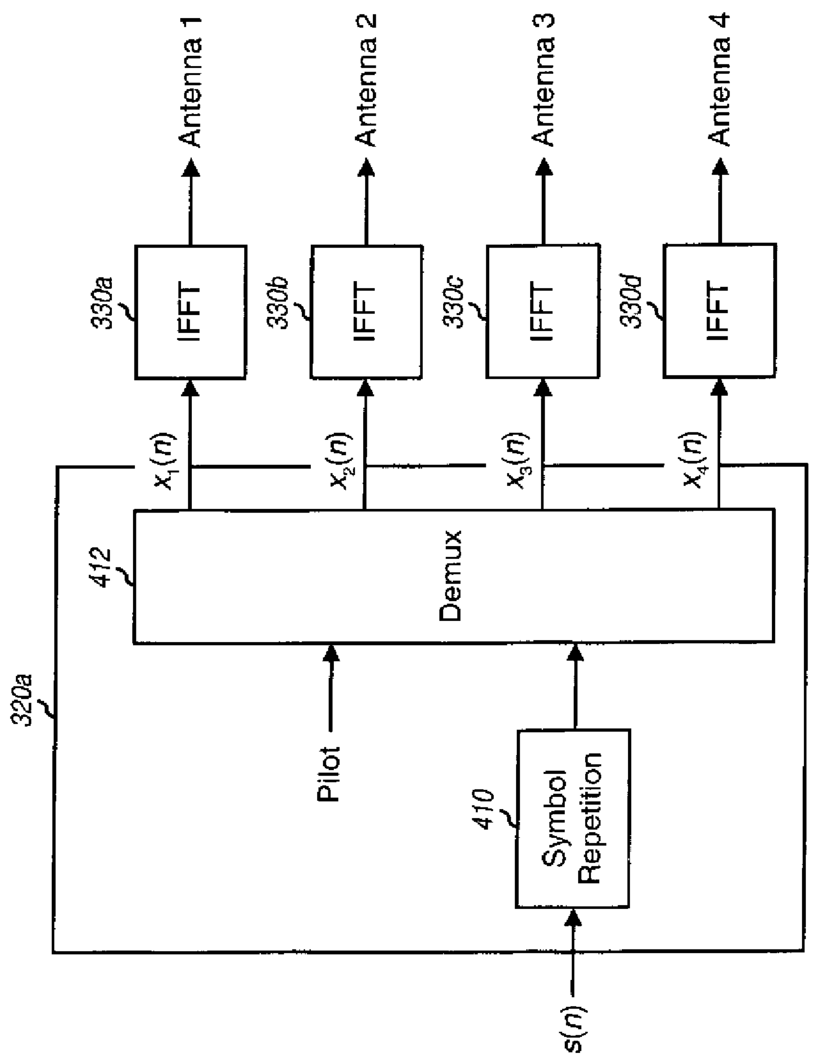

Antenna 1
Antenna 2
Antenna 3
Antenna 4
330a
330b
330c
330d
IFFT
IFFT
IFFT
IFFT
$x_1(n)$
$x_2(n)$
$x_3(n)$
$x_4(n)$
412
Demux
320a
Pilot
410
Symbol Repetition
$s(n)$

FIG. 4

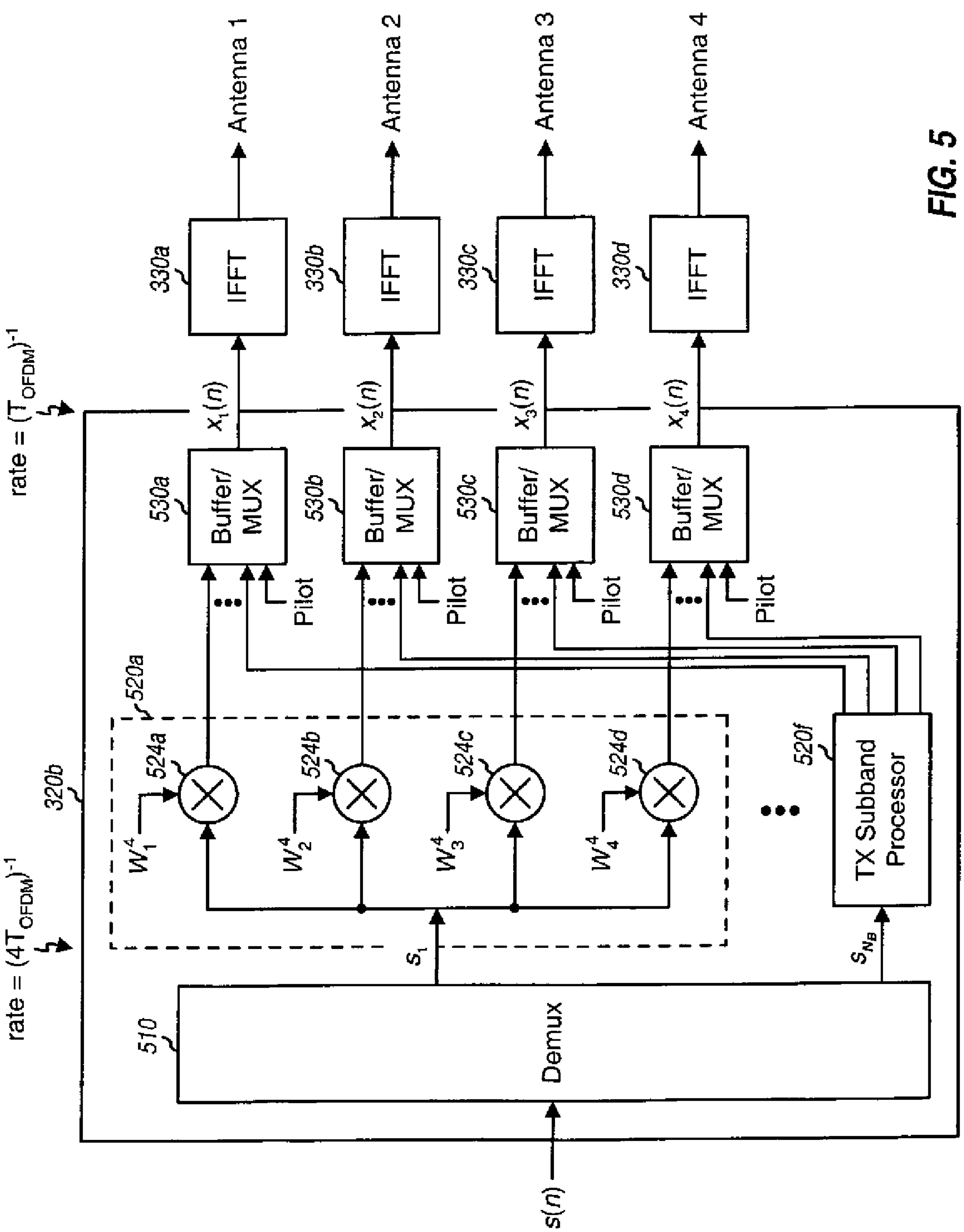

rate = $(4T_{OFDM})^{-1}$
rate = $(T_{OFDM})^{-1}$
320b
510
Demux
$s(n)$
s_1
s_{N_B}
520a
524a
524b
524c
524d
W_1^4
W_2^4
W_3^4
W_4^4
520f
TX Subband Processor
530a
530b
530c
530d
Buffer/ MUX
Pilot
$x_1(n)$
$x_2(n)$
$x_3(n)$
$x_4(n)$
330a
330b
330c
330d
IFFT
Antenna 1
Antenna 2
Antenna 3
Antenna 4

FIG. 5

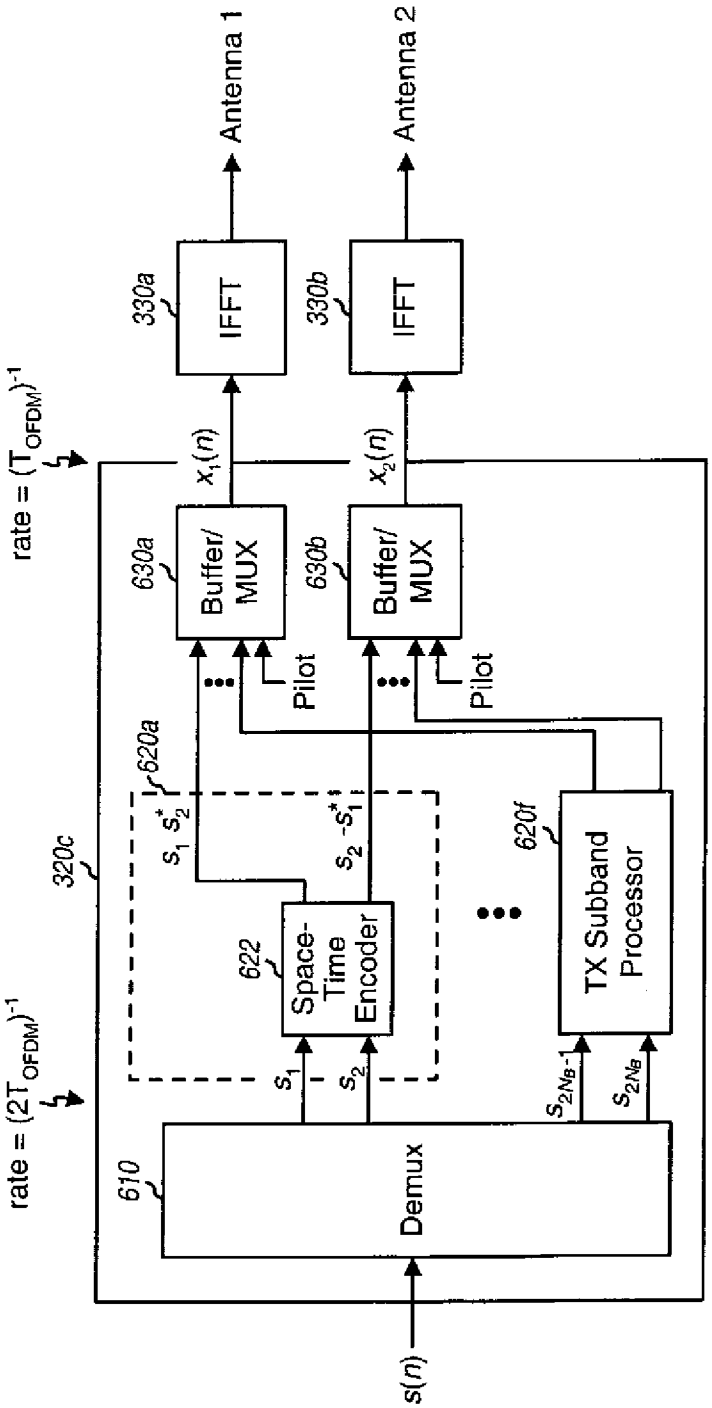

rate = (2T_OFDM)^-1
rate = (T_OFDM)^-1
320c
610
Demux
s(n)
s_1
s_2
s_2NB-1
s_2NB
620a
622
Space-Time Encoder
s_1 s_2*
s_2 -s_1*
620f
TX Subband Processor
Pilot
Pilot
630a
Buffer/MUX
630b
Buffer/MUX
x_1(n)
x_2(n)
330a
IFFT
330b
IFFT
Antenna 1
Antenna 2

FIG. 6

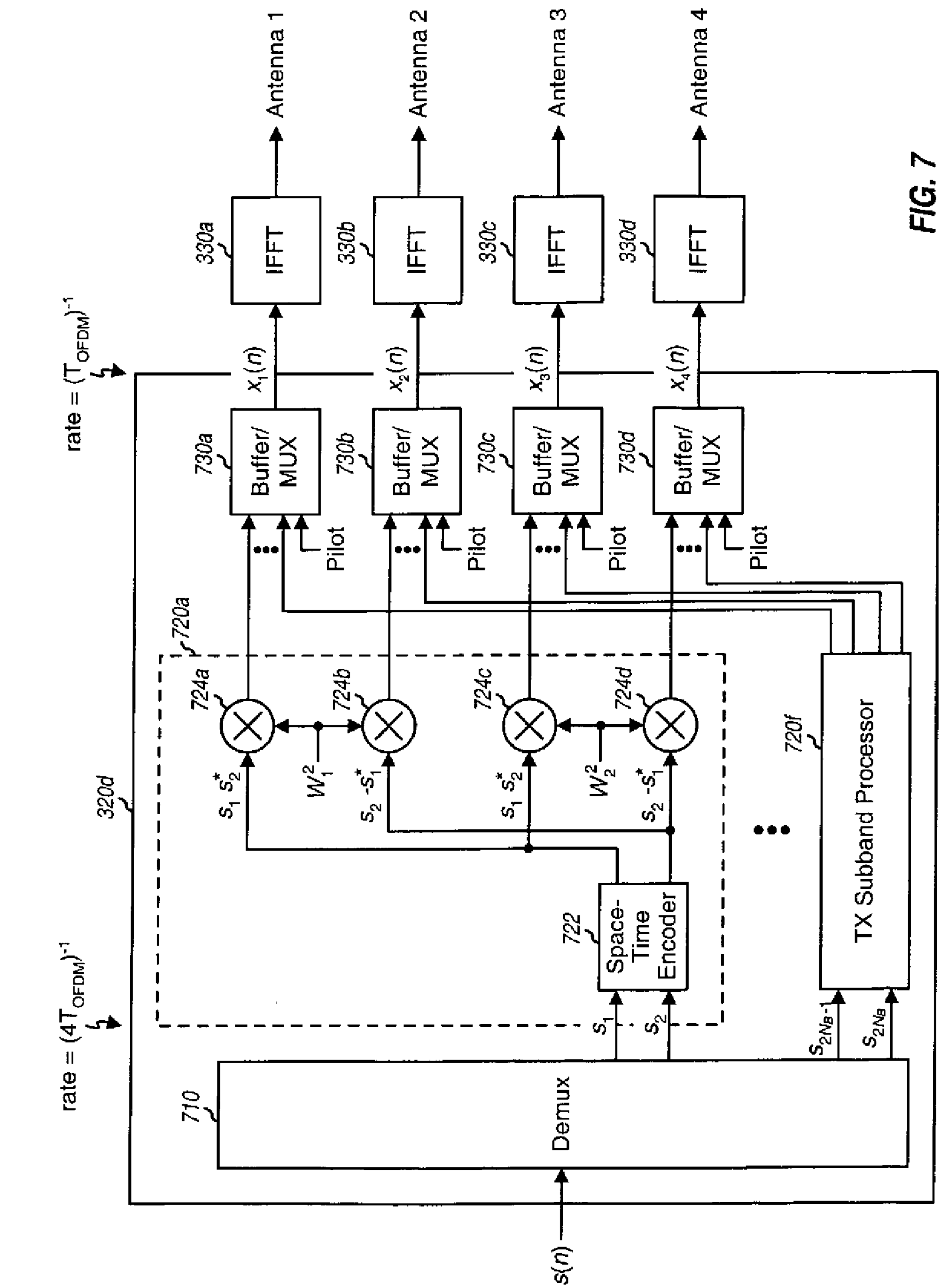
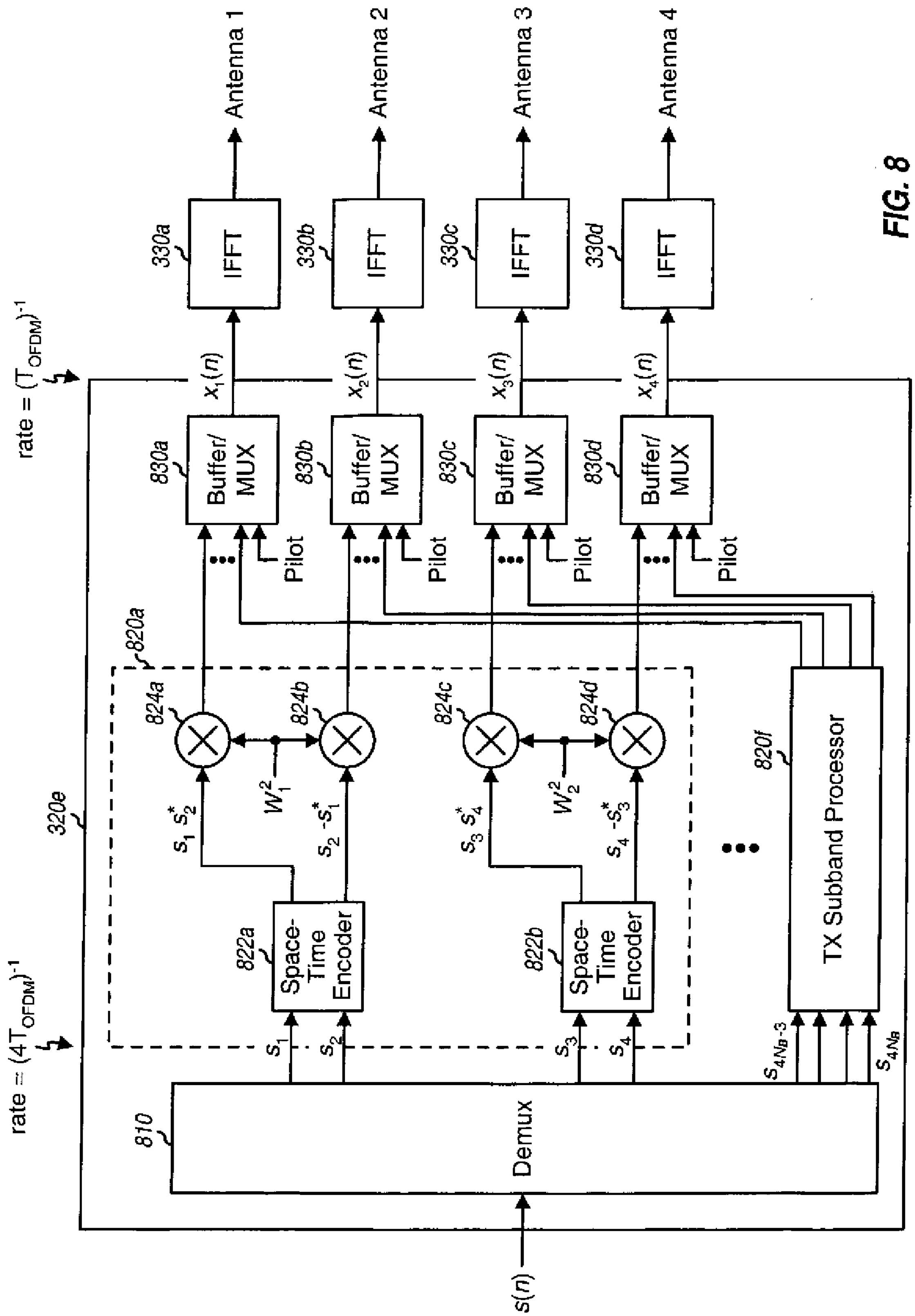

rate = $(4T_{OFDM})^{-1}$
rate = $(T_{OFDM})^{-1}$
320d
$s(n)$
710
Demux
s_1
s_2
720a
722
Space-Time Encoder
$s_1\ s_2^*$
$s_2\ {-s_1^*}$
W_1^2
W_2^2
724a
724b
724c
724d
s_{2N_B-1}
s_{2N_B}
720f
TX Subband Processor
Pilot
730a
730b
730c
730d
Buffer/ MUX
$x_1(n)$
$x_2(n)$
$x_3(n)$
$x_4(n)$
330a
330b
330c
330d
IFFT
Antenna 1
Antenna 2
Antenna 3
Antenna 4

FIG. 7



rate = $(4T_{OFDM})^{-1}$
rate = $(T_{OFDM})^{-1}$
320e
810
Demux
$s(n)$
s_1
s_2
s_3
s_4
s_{4N_B-3}
s_{4N_B}
820a
822a
Space-Time Encoder
822b
Space-Time Encoder
$s_1\ s_2^*$
$s_2\ -s_1^*$
$s_3\ s_4^*$
$s_4\ -s_3^*$
W_1^2
W_2^2
824a
824b
824c
824d
820f
TX Subband Processor
Pilot
830a
830b
830c
830d
Buffer/ MUX
$x_1(n)$
$x_2(n)$
$x_3(n)$
$x_4(n)$
330a
330b
330c
330d
IFFT
Antenna 1
Antenna 2
Antenna 3
Antenna 4

FIG. 8

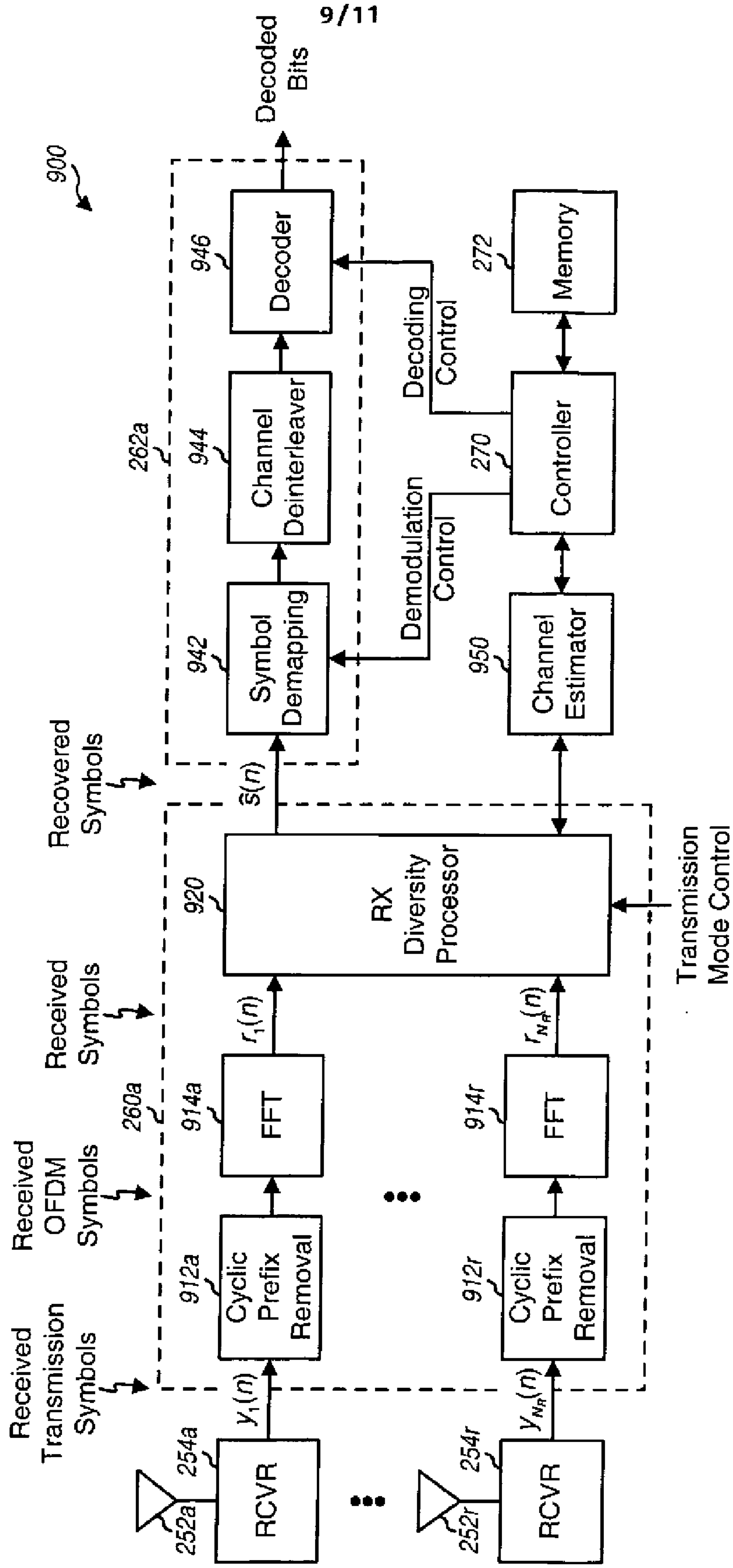

900
Decoded Bits
262a
946
Decoder
944
Channel Deinterleaver
942
Symbol Demapping
Decoding Control
Demodulation Control
272
Memory
270
Controller
950
Channel Estimator
Recovered Symbols
$\hat{s}(n)$
920
RX Diversity Processor
Transmission Mode Control
Received Symbols
$r_1(n)$
$r_{N_R}(n)$
260a
914a
FFT
914r
FFT
Received OFDM Symbols
912a
Cyclic Prefix Removal
912r
Cyclic Prefix Removal
Received Transmission Symbols
$y_1(n)$
$y_{N_R}(n)$
254a
RCVR
252a
254r
RCVR
252r

FIG. 9

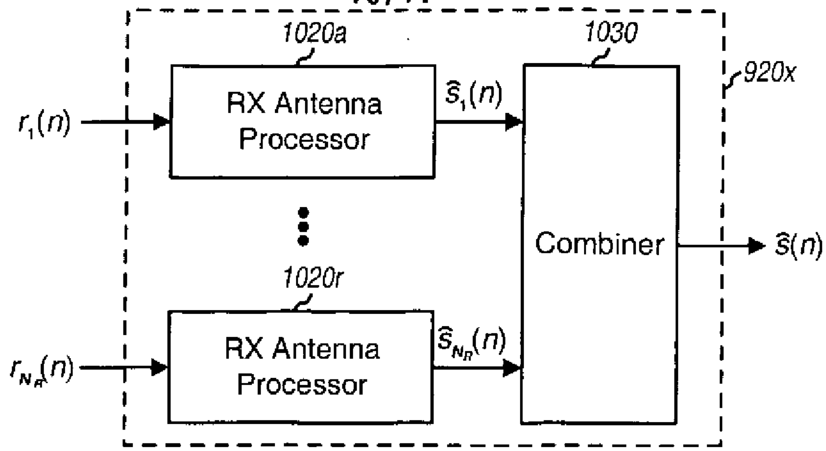

1020a
1030
920x
$r_1(n)$
RX Antenna Processor
$\hat{s}_1(n)$
Combiner
$\hat{s}(n)$
1020r
$r_{N_R}(n)$
RX Antenna Processor
$\hat{s}_{N_R}(n)$

FIG. 10

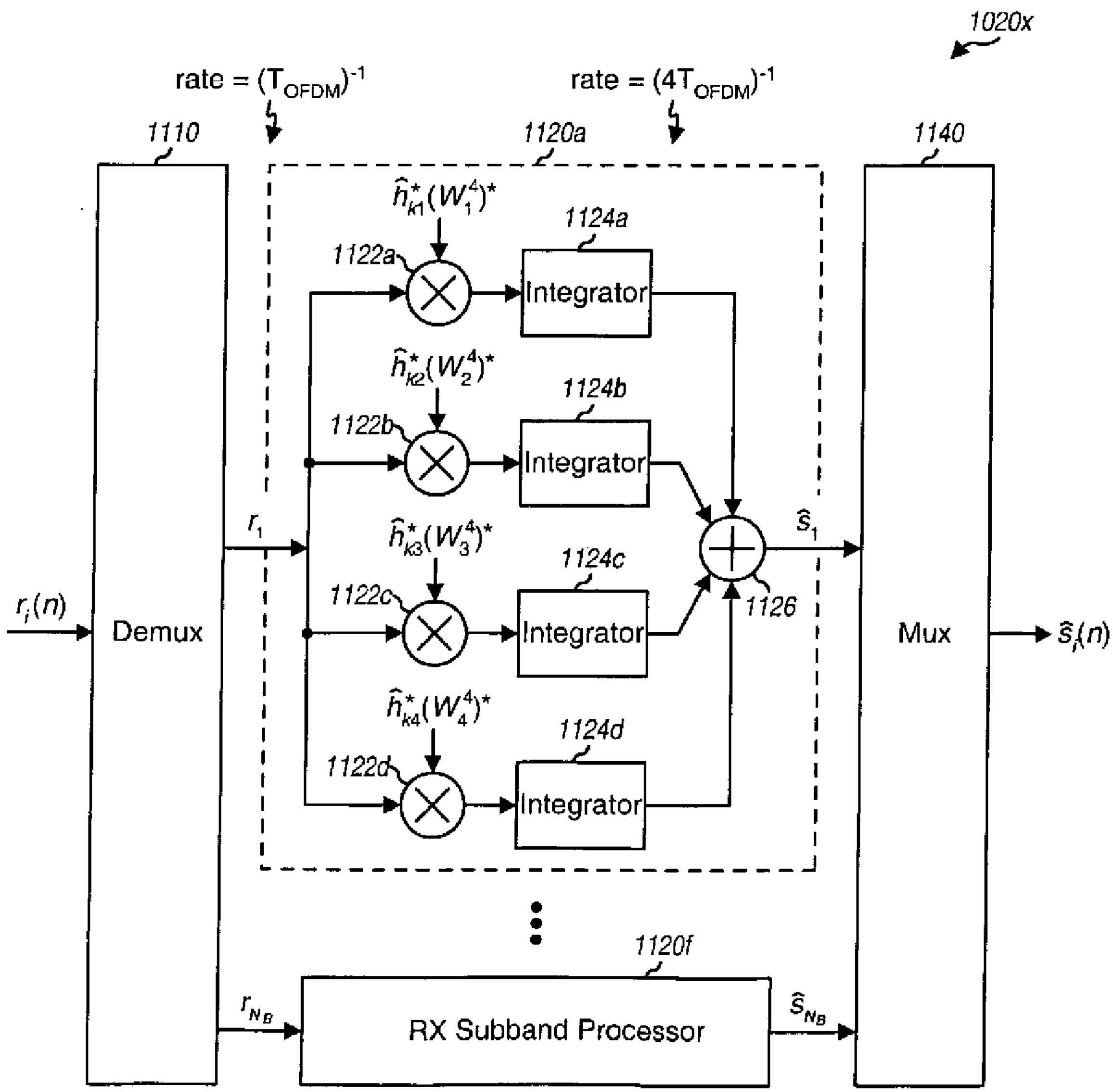

1020x
rate = $(T_{OFDM})^{-1}$
rate = $(4T_{OFDM})^{-1}$
1110
1120a
1140
$\hat{h}^*_{k1}(W^4_1)^*$
1124a
1122a
Integrator
$\hat{h}^*_{k2}(W^4_2)^*$
1124b
1122b
Integrator
r_1
$\hat{h}^*_{k3}(W^4_3)^*$
$\hat{s}_1$
1124c
$r_i(n)$
1122c
Demux
Integrator
1126
Mux
$\hat{s}_i(n)$
$\hat{h}^*_{k4}(W^4_4)^*$
1124d
1122d
Integrator
1120f
r_{N_B}
$\hat{s}_{N_B}$
RX Subband Processor

FIG. 11

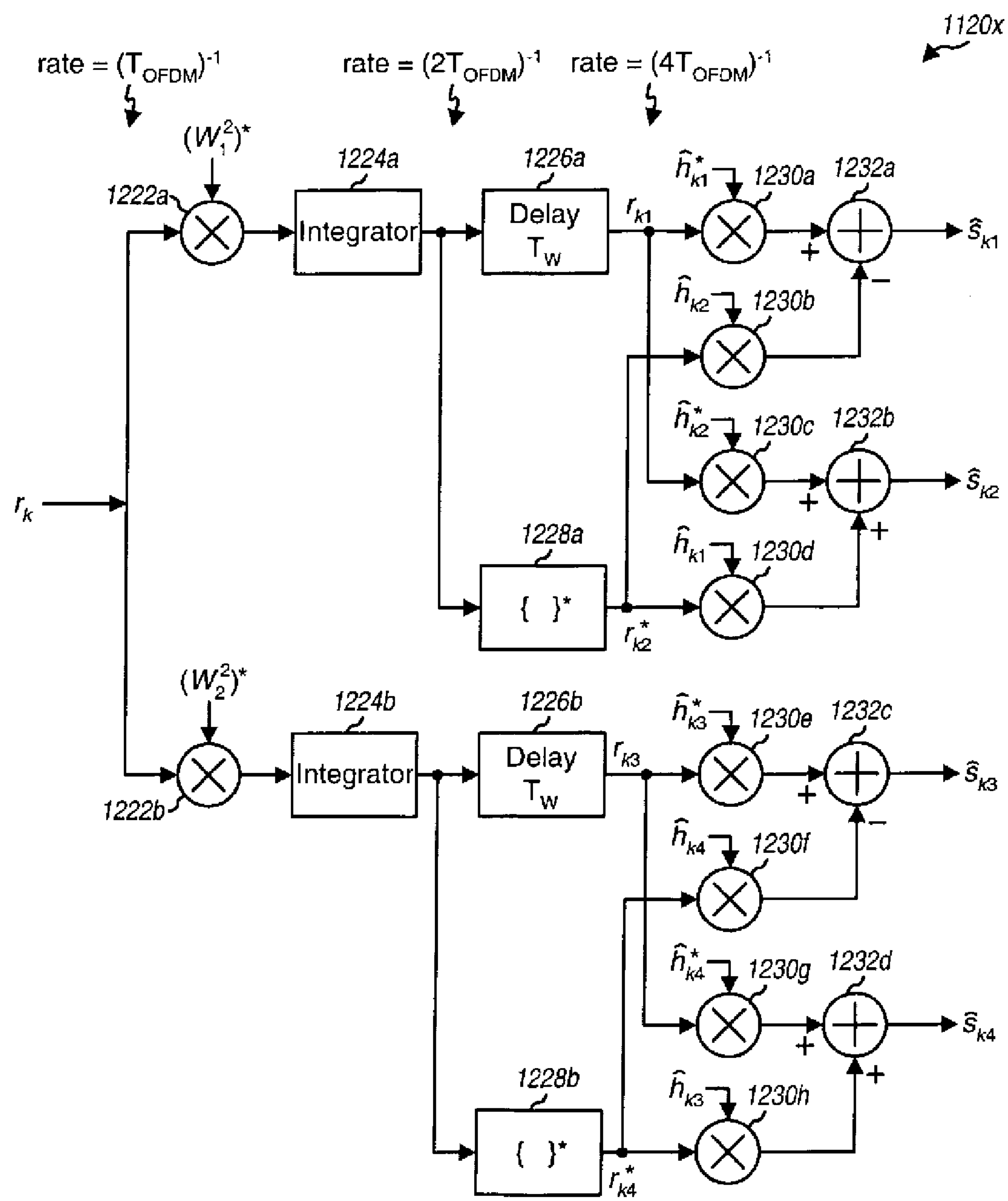

1120x
rate = $(T_{OFDM})^{-1}$
rate = $(2T_{OFDM})^{-1}$
rate = $(4T_{OFDM})^{-1}$
$(W_1^2)^*$
$(W_2^2)^*$
1222a
1222b
1224a
1224b
Integrator
Integrator
1226a
1226b
Delay T_W
Delay T_W
1228a
1228b
{ }*
{ }*
r_k
r_{k1}
r_{k2}^*
r_{k3}
r_{k4}^*
$\hat{h}_{k1}^*$
$\hat{h}_{k2}$
$\hat{h}_{k2}^*$
$\hat{h}_{k1}$
$\hat{h}_{k3}^*$
$\hat{h}_{k4}$
$\hat{h}_{k4}^*$
$\hat{h}_{k3}$
1230a
1230b
1230c
1230d
1230e
1230f
1230g
1230h
1232a
1232b
1232c
1232d
$\hat{s}_{k1}$
$\hat{s}_{k2}$
$\hat{s}_{k3}$
$\hat{s}_{k4}$

FIG. 12

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 03/19466

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04B7/04 H04L27/26 H04L1/06 H04B7/06 H04J11/00
H04L1/00

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04B H04L H04J

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	WO 01 76110 A (QUALCOMM INC) 11 October 2001 (2001-10-11)	1,2, 11-21, 24,28, 32-34, 38-42, 44,45, 47-49
Y		3-10,22, 23, 25-27, 29-31, 35-37, 43,46
	page 8, line 24 -page 9, line 21 page 12, line 5 -page 13, line 11 page 18, line 21 -page 19, line 13 page 27, line 18 -page 28, line 5 page 32, line 22 - line 29; figure 3 claims 1,8-10	
	-/--	

[X] Further documents are listed in the continuation of box C.

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Date of the actual completion of the international search	Date of mailing of the international search report
28 October 2003	04/11/2003

Name and mailing address of the ISA	Authorized officer
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INTERNATIONAL SEARCH REPORT

Internati ›plication No
PCT/US 03/19466

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
Y	EP 1 182 799 A (LUCENT TECHNOLOGIES INC) 27 February 2002 (2002-02-27) paragraph '0024! paragraph '0048!	3-10,22, 23, 25-27, 29-31, 35-37, 43,46
A	EP 1 158 716 A (AT & T CORP) 28 November 2001 (2001-11-28) paragraph '0009! paragraph '0014! - paragraph '0016!; figure 4	1-49

Form PCT/ISA/210 (continuation of second sheet) (July 1992)

INTERNATIONAL SEARCH REPORT

Information on patent family members

Internatio   pplication No
PCT/US 03/19466

Patent document cited in search report		Publication date	Patent family member(s)			Publication date
WO 0176110	A	11-10-2001	US	6473467	B1	29-10-2002
			AU	4592101	A	15-10-2001
			CA	2402152	A1	11-10-2001
			EP	1269665	A2	02-01-2003
			NO	20024634	A	20-11-2002
			TW	508922	B	01-11-2002
			WO	0176110	A2	11-10-2001
EP 1182799	A	27-02-2002	BR	0103361	A	23-04-2002
			CA	2351844	A1	22-02-2002
			CN	1339885	A	13-03-2002
			EP	1182799	A2	27-02-2002
			JP	2002135232	A	10-05-2002
EP 1158716	A	28-11-2001	US	2001053143	A1	20-12-2001
			CN	1325198	A	05-12-2001
			EP	1158716	A2	28-11-2001
			JP	2002044051	A	08-02-2002
			TW	510103	B	11-11-2002

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INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(51) International Patent Classification 6 : H04B 7/26, 1/707	A1	(11) International Publication Number: WO 96/35268 (43) International Publication Date: 7 November 1996 (07.11.96)

(21) International Application Number: PCT/US96/07567

(22) International Filing Date: 2 May 1996 (02.05.96)

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436,029 5 May 1995 (05.05.95) US

(71) Applicant: QUALCOMM INCORPORATED [US/US]; 6455 Lusk Boulevard, San Diego, CA 92121 (US).

(72) Inventors: ZIV, Noam, A.; 10968 Corte Playa Barcelona, San Diego, CA 92124 (US). PADOVANI, Roberto; 13593 Penfield Point, San Diego, CA 92130 (US). LEVIN, Jeffrey, A.; 12549 Maestro Court, San Diego, CA 92130 (US). EASTON, Kenneth, D.; 7379 Calle Cristobal #217, San Diego, CA 92126 (US).

(74) Agent: MILLER, Russell, B.; Qualcomm Incorporated, 6455 Lusk Boulevard, San Diego, CA 92121 (US).

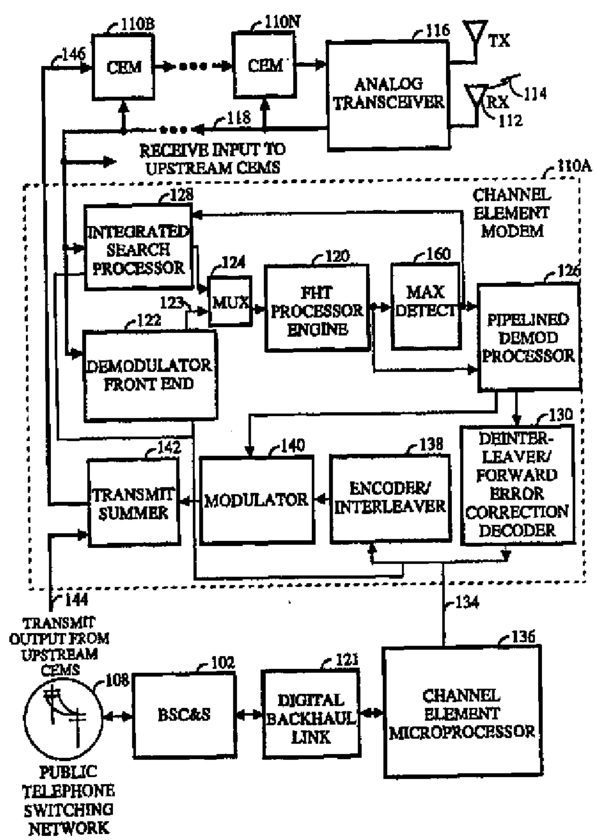
(81) Designated States: AL, AM, AT, AU, AZ, BB, BG, BR, BY, CA, CH, CN, CZ, DE, DK, EE, ES, FI, GB, GE, HU, IS, JP, KE, KG, KP, KR, KZ, LK, LR, LS, LT, LU, LV, MD, MG, MK, MN, MW, MX, NO, NZ, PL, PT, RO, RU, SD, SE, SG, SI, SK, TJ, TM, TR, TT, UA, UG, UZ, VN, ARIPO patent (KE, LS, MW, SD, SZ, UG), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, CH, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, ML, MR, NE, SN, TD, TG).

Published
With international search report.
Before the expiration of the time limit for amending the claims and to be republished in the event of the receipt of amendments.

(54) Title: METHOD OF RECEIVING AND SEARCHING A SIGNAL TRANSMITTING IN BURSTS

(57) Abstract

An integrated search processor (128) used in a modem (110) for a spread spectrum communications system buffers in a buffer (172) received signals samples and utilizes a time sliced transform processor (120) operating on successive offsets from the buffer (172). The search processor (128) autonomously steps through a search as configured by a microprocessor (136) specified search parameter set, which can include the group of antennas to search over, the starting offset and width of the search window to search over, and the number of Walsh symbols to accumulate results at each offset. The search processor (128) calculates the correlation energy at each offset, and presents a summary report of the best paths found in the search to use for demodulation element (122) reassignment. The search is done in a linar fashion independent of the probability that a signal being searched for was transmitted at any given time.



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METHOD OF RECEIVING AND SEARCHING A SIGNAL TRANSMITTED IN BURSTS

BACKGROUND OF THE INVENTION

I. Field of the Invention

The present application is a continuation-in-part application of copending U.S. Patent Application Serial No. 08/316,177, filed September 30, 1994, entitled MULTIPATH SEARCH PROCESSOR FOR A SPREAD SPECTRUM MULTIPLE ACCESS COMMUNICATION SYSTEM. The present invention relates generally to spread spectrum communication systems and, more particularly, to signal processing in a cellular telephone communication system.

II. Description of the Related Art

In a wireless telephone communication system such as cellular telephone systems, personal communications systems, and wireless local loop system, many users communicate over a wireless channel to connect to wireline telephone systems. Communication over the wireless channel can be one of a variety of multiple access techniques which facilitate a large number of users in a limited frequency spectrum. These multiple access techniques include time division multiple access (TDMA), frequency division multiple access (FDMA), and code division multiple access (CDMA). The CDMA technique has many advantages and an exemplary CDMA system is described in U.S. Patent No. 4,901,307 issued February 13, 1990 to K. Gilhousen et al., entitled "SPREAD SPECTRUM MULTIPLE ACCESS COMMUNICATION SYSTEM USING SATELLITE OR TERRESTRIAL REPEATERS," assigned to the assignee of the present invention and incorporated herein by reference.

In the just mentioned patent, a multiple access technique is disclosed where a large number of mobile telephone system users, each having a transceiver, communicate through satellite repeaters or terrestrial base stations using CDMA spread spectrum communication signals. In using CDMA communications, the frequency spectrum can be reused multiple times thus permitting an increase in system user capacity.

The CDMA modulation techniques disclosed in U.S. Patent No. 4,901,307 offer many advantages over narrow band modulation techniques used in communication systems using satellite or terrestrial channels. The terrestrial channel poses special problems to any

communication system particularly with respect to multipath signals. The use of CDMA techniques permits the special problems of the terrestrial channel to be overcome by mitigating the adverse effect of multipath, e.g. fading, while also exploiting the advantages thereof.

The CDMA techniques as disclosed in U.S. Patent No. 4,901,307 contemplate the use of coherent modulation and demodulation for both directions of the link in remote unit-satellite communications. Accordingly, disclosed therein is the use of a pilot carrier signal as a coherent phase reference for the satellite-to-remote unit link and the base station-to-remote unit link. In the terrestrial cellular environment, however, the severity of multipath fading with the resulting phase disruption of the channel, as well as the power required to transmit a pilot carrier signal from the remote unit, precludes usage of coherent demodulation techniques for the remote unit-to-base station link. U.S. Patent No. 5,103,459 entitled "SYSTEM AND METHOD FOR GENERATING SIGNAL WAVEFORMS IN A CDMA CELLULAR TELEPHONE SYSTEM", issued June 25, 1990, assigned to the assignee of the present invention, the disclosure of which is incorporated by this reference, provides a means for overcoming the adverse effects of multipath in the remote unit-to-base station link by using noncoherent modulation and demodulation techniques.

In a CDMA cellular telephone system, the same frequency band can be used for communication in all base stations. At the base station receiver, separable multipath, such as a line of site path and another path reflecting off of a building, can be diversity combined for enhanced modem performance. The CDMA waveform properties that provide processing gain are also used to discriminate between signals that occupy the same frequency band. Furthermore, the high speed pseudonoise (PN) modulation allows many different propagation paths of the same signal to be separated, provided the difference in path delays exceeds the PN chip duration. If a PN chip rate of approximately 1 MHz is employed in a CDMA system, the full spread spectrum processing gain, equal to the ratio of the spread bandwidth to the system data rate, can be employed against paths having delays that differ by more than one microsecond. A one microsecond path delay differential corresponds to differential path distance of approximately 300 meters. The urban environment typically provides differential path delays in excess of one microsecond.

The multipath properties of the terrestrial channel produce at the receiver signals having traveled several distinct propagation paths. One characteristic of a multipath channel is the time spread introduced in a signal

that is transmitted through the channel. For example, if an ideal impulse is transmitted over a multipath channel, the received signal appears as a stream of pulses. Another characteristic of the multipath channel is that each path through the channel may cause a different attenuation factor. For example, if an ideal impulse is transmitted over a multipath channel, each pulse of the received stream of pulses generally has a different signal strength than other received pulses. Yet another characteristic of the multipath channel is that each path through the channel may cause a different phase on the signal. For example, if an ideal impulse is transmitted over a multipath channel, each pulse of the received stream of pulses generally has a different phase than other received pulses.

In the radio channel, the multipath is created by reflection of the signal from obstacles in the environment, such as buildings, trees, cars, and people. In general the radio channel is a time varying multipath channel due to the relative motion of the structures that create the multipath. For example, if an ideal impulse is transmitted over the time varying multipath channel, the received stream of pulses would change in time location, attenuation, and phase as a function of the time that the ideal impulse was transmitted.

The multipath characteristic of a channel can result in signal fading. Fading is the result of the phasing characteristics of the multipath channel. A fade occurs when multipath vectors are added destructively, yielding a received signal that is smaller than either individual vector. For example, if a sine wave is transmitted through a multipath channel having two paths where the first path has an attenuation factor of X dB, a time delay of δ with a phase shift of Θ radians, and the second path has an attenuation factor of X dB, a time delay of δ with a phase shift of $\Theta + \pi$ radians, no signal would be received at the output of the channel.

In narrow band modulation systems such as the analog FM modulation employed by conventional radio telephone systems, the existence of multiple paths in the radio channel results in severe multipath fading. As noted above with a wideband CDMA, however, the different paths may be discriminated in the demodulation process. This discrimination not only greatly reduces the severity of multipath fading but provides an advantage to the CDMA system.

Diversity is one approach for mitigating the deleterious effects of fading. It is therefore desirable that some form of diversity be provided which permits a system to reduce fading. Three major types of diversity exist: time diversity, frequency diversity, and space/path diversity.

Time diversity can best be obtained by the use of repetition, time interleaving, and error correction and detection coding which introduce redundancy. A system comprising the present invention may employ each of these techniques as a form of time diversity.

CDMA by its inherent wideband nature offers a form of frequency diversity by spreading the signal energy over a wide bandwidth. Therefore, frequency selective fading affects only a small part of the CDMA signal bandwidth.

Space and path diversity are obtained by providing multiple signal paths through simultaneous links from a remote unit through two or more base stations and by employing two or more spaced apart antenna elements at a single base station. Furthermore, path diversity may be obtained by exploiting the multipath environment through spread spectrum processing by allowing a signal arriving with different propagation delays to be received and processed separately as discussed above. Examples of path diversity are illustrated in U.S. Patent No. 5,101,501 entitled "SOFT HANDOFF IN A CDMA CELLULAR TELEPHONE SYSTEM", issued March 21, 1992 and U.S. Patent No. 5,109,390 entitled "DIVERSITY RECEIVER IN A CDMA CELLULAR TELEPHONE SYSTEM", issued April 28, 1992, both assigned to the assignee of the present invention.

The deleterious effects of fading can be further controlled to a certain extent in a CDMA system by controlling transmitter power. A system for base station and remote unit power control is disclosed in U.S. Patent No. 5,056,109 entitled "METHOD AND APPARATUS FOR CONTROLLING TRANSMISSION POWER IN A CDMA CELLULAR MOBILE TELEPHONE SYSTEM", issued October 8, 1991, also assigned to the assignee of the present invention.

The CDMA techniques as disclosed in U.S. Patent No. 4,901,307 contemplate the use of relatively long PN sequences with each remote unit user being assigned a different PN sequence. The cross-correlation between different PN sequences and the autocorrelation of a PN sequence, for all time shifts other than zero, both have a nearly zero average value which allows the different user signals to be discriminated upon reception. (Autocorrelation and cross-correlation requires logical "0" take on a value of "1" and logical "1" take on a value of "-1" or a similar mapping in order that a zero average value be obtained.)

However, such PN signals are not orthogonal. Although the cross-correlation essentially averages to zero over the entire sequence length, for a short time interval, such as an information bit time, the cross-correlation is a

random variable with a binomial distribution. As such, the signals interfere with each other in much the same way as they would if they were wide bandwidth Gaussian noise at the same power spectral density. Thus the other user signals, or mutual interference noise, ultimately limits the achievable capacity.

It is well known in the art that a set of n orthogonal binary sequences, each of length n, for n any power of 2 can be constructed, see Digital Communications with Space Applications, S.W. Golomb et al., Prentice-Hall, Inc., 1964, pp. 45-64. In fact, orthogonal binary sequence sets are also known for most lengths which are multiples of four and less than two hundred. One class of such sequences that is easy to generate is called the Walsh function, also known as Hadamard matrices.

A Walsh function of order n can be defined recursively as follows:

$$W(n) = \begin{vmatrix} W(n/2)\ , W(n/2) \\ W(n/2)\ , W'(n/2) \end{vmatrix}$$

where W' denotes the logical complement of W, and $W(1) = \begin{vmatrix} 0 \end{vmatrix}$.

Thus,

$$W(2) = \begin{vmatrix} 0, & 0 \\ 0, & 1 \end{vmatrix},$$

$$W(4) = \begin{vmatrix} 0, & 0, & 0, & 0 \\ 0, & 1, & 0, & 1 \\ 0, & 0, & 1, & 1 \\ 0, & 1, & 1, & 0 \end{vmatrix}, \text{ and}$$

$$W(8) = \begin{vmatrix} 0, & 0, & 0, & 0, & 0, & 0, & 0, & 0 \\ 0, & 1, & 0, & 1, & 0, & 1, & 0, & 1 \\ 0, & 0, & 1, & 1, & 0, & 0, & 1, & 1 \\ 0, & 1, & 1, & 0, & 0, & 1, & 1, & 0 \\ 0, & 0, & 0, & 0, & 1, & 1, & 1, & 1 \\ 0, & 1, & 0, & 1, & 1, & 0, & 1, & 0 \\ 0, & 0, & 1, & 1, & 1, & 1, & 0, & 0 \\ 0, & 1, & 1, & 0, & 1, & 0, & 0, & 1 \end{vmatrix}.$$

A Walsh symbol, sequence, or code is one of the rows of a Walsh function matrix. A Walsh function matrix of order n contains n sequences, each of length n Walsh chips. Each Walsh code has a corresponding Walsh index where the Walsh index refers to the number (1 through n) corresponding to the row in which a Walsh code is found. For example, for n=8 Walsh function matrix given above, the all zeroes row corresponds to

Walsh index 1 and the Walsh code 0, 0, 0, 0, 1, 1, 1, 1 corresponds to Walsh index 5.

A Walsh function matrix of order n (as well as other orthogonal functions of length n) has the property that over the interval of n bits, the cross-correlation between all the different sequences within the set is zero. This can be seen by noting that every sequence differs from every other sequence in exactly half of its bits. It should also be noted that there is always one sequence containing all zeroes and that all the other sequences contain half ones and half zeroes. The Walsh symbol which consists all logical zeros instead of half one's and zero's is called the Walsh zero symbol.

On the reverse link channel from the remote unit to the base station, no pilot signal exists to provide a phase reference. Therefore a method is needed to provide a high-quality link on a fading channel having a low Eb/No (energy per bit/noise power density). Walsh function modulation on the reverse link is a simple method of obtaining 64-ary modulation with coherence over the set of six code symbols mapped into the 64 Walsh codes. The characteristics of the terrestrial channel are such that the rate of change of phase is relatively slow. Therefore, by selecting a Walsh code duration which is short compared to the rate of change of phase on the channel, coherent demodulation over the length of one Walsh code is possible.

On the reverse link channel, the Walsh code is determined by the information being transmitted from the remote unit. For example, a three bit information symbol could be mapped into the eight sequences of W(8) given above. An "unmapping" of the Walsh encoded symbols into an estimate of the original information symbols may be accomplished in the receiver by a Fast Hadamard Transform (FHT). A preferred "unmapping" or selection process produces soft decision data which can be provided to a decoder for maximum likelihood decoding.

An FHT is used to perform the "unmapping" process. An FHT correlates the received sequence with each of the possible Walsh sequences. Selection circuitry is employed to select the most likely correlation value, which is scaled and provided as soft decision data.

A spread spectrum receiver of the diversity or "rake" receiver design comprises multiple data receivers to mitigate the effects of fading. Typically each data receiver is assigned to demodulate a signal which has traveled a different path, either through the use of multiple antennas or due to the multipath properties of the channel. In the demodulation of signals modulated according to an orthogonal signaling scheme, each data receiver correlates the received signal with each of the possible mapping values using

an FHT. The FHT outputs of each data receiver are combined and selection circuitry then selects the most likely correlation value based on the largest combined FHT output to produce a demodulated soft decision symbol.

In the system described in the U.S. Patent No. 5,103,459, the call signal begins as a 9600 bit per second information source which is then converted by a rate 1/3 forward error correction encoder to a 28,800 symbols per second output stream. These symbols are grouped 6 at a time to form 4800 Walsh symbols per second, each Walsh symbol selecting one of sixty-four orthogonal Walsh functions that are sixty-four Walsh chips in duration. The Walsh chips are modulated with a user-specific PN sequence generator. The user-specific PN modulated data is then split into two signals, one of which is modulated with an in-phase (I) channel PN sequence and one of which is modulated with a quadrature-phase (Q) channel PN sequence. Both the I channel modulation and the Q channel modulation provide four PN chips per Walsh chip with a 1.2288 MHz PN spreading rate. The I and the Q modulated data are Offset Quadrature Phase Shift Keying (OQPSK) combined for transmission.

In the CDMA cellular system described in the above-referenced U.S. Patent No. 4,901,307, each base station provides coverage to a limited geographic area and links the remote units in its coverage area through a cellular system switch to the public switched telephone network (PSTN). When a remote unit moves to the coverage area of a new base station, the routing of that user's call is transferred to the new base station. The base station-to-remote unit signal transmission path is referred to as the forward link and the remote unit-to-base station signal transmission path is referred to as the reverse link.

As described above, the PN chip interval defines the minimum separation two paths must have in order to be combined. Before the distinct paths can be demodulated, the relative arrival times (or offsets) of the paths in the received signal must first be determined. The channel element modem performs this function by "searching" through a sequence of potential path offsets and measuring the energy received at each potential path offset. If the energy associated with a potential offset exceeds a certain threshold, a signal demodulation element may be assigned to that offset. The signal present at that path offset can then be summed with the contributions of other demodulation elements at their respective offsets. A method and apparatus of demodulation element assignment based on searcher demodulation element energy levels is disclosed in co-pending U.S. Patent Application Serial No. 08/144,902 entitled "DEMODULATION ELEMENT ASSIGNMENT

IN A SYSTEM CAPABLE OF RECEIVING MULTIPLE SIGNALS," filed October 28, 1993, assigned to the assignee of the present invention. Such a diversity or rake receiver provides for a robust digital link, because all paths have to fade together before the combined signal is degraded.

FIG. 1 shows an exemplary set of signals from a single remote unit arriving at the base station. The vertical axis represents the power received on a decibel (dB) scale. The horizontal axis represents the delay in the arrival time of a signal due to multipath delays. The axis (not shown) going into the page represents a segment of time. Each signal spike in the common plane of the page has arrived at a common time but was transmitted by the remote unit at a different time. In a common plane, peaks to the right were transmitted at an earlier time by the remote unit than peaks to the left. For example, the left-most peak spike **2** corresponds to the most recently transmitted signal. Each signal spike **2 - 7** has traveled a different path and therefore exhibits a different time delay and a different amplitude response. The six different signal spikes represented by spikes **2 - 7** are representative of a severe multipath environment. Typical urban environments produce fewer usable paths. The noise floor of the system is represented by the peaks and dips having lower energy levels. The task of a searcher element is to identify the delay as measured by the horizontal axis of signal spikes **2 - 7** for potential demodulation element assignment. The task of the demodulation element is to demodulate a set of the multipath peaks for combination into a single output. It is also the task of the demodulation elements once assigned to a multipath peak to track that peak as it may move in time.

The horizontal axis can also be thought of as having units of PN offset. At any given time, the base station receives a variety of signals from a single remote unit, each of which has traveled a different path and may have a different delay than the others. The remote unit's signal is modulated by a PN sequence. A copy of the PN sequence is also generated at the base station. At the base station, each multipath signal is individually demodulated with a PN sequence code aligned to its timing. The horizontal axis coordinates can be thought of as corresponding to the PN sequence code offset which would be used to demodulate a signal at that coordinate.

Note that each of the multipath peaks varies in amplitude as a function of time as shown by the uneven ridge of each multipath peak. In the limited time shown, there are no major changes in the multipath peaks. Over a more extended time range, multipath peaks disappear and new paths are created as time progresses. The peaks can also slide to earlier or later offsets as the path distances change as the remote unit moves around in the coverage area of the

base station. Each demodulation element tracks small variations in the signal assigned to it. The task of the searching process is to generate a log of the current multipath environment as received by the base station.

In a typical wireless telephone communication system, the remote unit transmitter may employ a vocoding system which encodes voice information in a variable rate format. For example, the data rate may be lowered due to pauses in the voice activity. The lower data rate reduces the level of interference to other users caused by the remote unit transmissions. At the receiver, or otherwise associated with the receiver, a vocoding system is employed for reconstructing the voice information. In addition to voice information, non-voice information alone or a mixture of the two may be transmitted by the remote unit.

A vocoder which is suited for application in this environment is described in copending U.S. patent application Serial No. 08/363,170, entitled "VARIABLE RATE VOCODER," filed December 23, 1994 and assigned to the assignee of the present invention. This vocoder produces from digital samples of the voice information encoded data at four different rates, e.g. approximately 8,000 bits per second (bps), 4,000 bps, 2,000 bps and 1,000 bps, based on voice activity during a 20 millisecond (ms) frame. Each frame of vocoder data is formatted with overhead bits as 9,600 bps, 4,800 bps, 2,400 bps, and 1,200 bps data frames. The highest rate data frame which corresponds to a 9,600 bps frame is referred to as a "full rate" frame; a 4,800 bps data frame is referred to as a "half rate" frame; a 2,400 bps data frame is referred to as a "quarter rate" frame; and a 1,200 bps data frame is referred to as an "eighth rate" frame. In neither the encoding process nor the frame formatting process is rate information included in the data. When the remote unit transmits data at less than full rate, the duty cycle of the remote units transmitted signal is the same as the data rate. For example, at quarter rate a signal is transmitted from the remote unit only one quarter of the time. During the other three quarters time, no signal is transmitted from the remote unit.

The remote unit includes a data burst randomizer. The data burst randomizer determines during which time periods the remote unit transmits and during which time periods it does not transmit given the data rate of the signal to be transmitted, a remote unit specific identifying number, and the time of day. When operating at less than full rate, the data burst randomizer within the remote unit pseudorandomly distributes the active time periods within the transmission burst. A corresponding data burst randomizer is also included in the base station such that the base station can recreate the pseudorandom distribution based on the time of day and the remote unit

specific identifying number but the base station is unaware, a priori, of the data rate of the transmitted signal.

The eighth rate time periods determine a so called worthy group of time periods. A remote unit operating at quarter rate transmits during the worthy group time periods and another set of pseudorandomly distributed periods. A remote unit operating in half rate transmits during the quarter rate time periods and another set of pseudorandomly distributed periods. A remote unit operating in full rate transmits continually. In this way, regardless of the data rate of the transmitted signal, each time period corresponding to the worthy group is sure to correspond to a time when the corresponding remote unit is transmitting a signal. Further details on the data burst randomizer are described in copending U.S. patent application Serial No. 08/291,647, entitled "DATA BURST RANDOMIZER," filed August 16, 1994, and assigned to the assignee of the present invention.

To conserve system resources for actual data of voice transmissions, the remote unit does not transmit the rate for each frame. Therefore, the receiver must determine the rate at which the data was encoded and transmitted based on the received signal so that the receiver associated vocoder can properly reconstruct the voice information. A method of determining the rate at which burst data was encoded without receiving rate information from the transmitter is disclosed in co-pending U.S. Patent Serial No. 08/233,570, entitled "METHOD AND APPARATUS FOR DETERMINING DATA RATE OF TRANSMITTED VARIABLE RATE DATE IN A COMMUNICATIONS RECEIVER" filed April 26, 1994, and assigned to the assignee of the present invention. The method of determining data rate disclosed in the above mentioned patent application is performed after the signal has been received and demodulated therefore the rate information is not available during the searching process.

At the base station, each individual remote unit signal must be identified from the ensemble of call signals received. A system and method for demodulating a remote unit signal received at a base station is described, for example, in U.S. Patent No. 5,103,459. FIG. 2 is a block diagram of the base station equipment described in U.S. Patent No. 5,103,459 for demodulating a reverse link remote unit signal.

A typical prior art base station comprises multiple independent searcher and demodulation elements. The searcher and demodulation elements are controlled by a microprocessor. In this exemplary embodiment, to maintain a high system capacity, each remote unit in the system does not transmit a pilot signal. The lack of a pilot signal on the reverse link increases

the time needed to conduct a survey of all possible time offsets at which a remote unit signal may be received. Typically, a pilot signal is transmitted at a higher power than the traffic bearing signals thus increasing the signal to noise ratio of the received pilot signal as compared to the received traffic channel signals. In contrast, ideally each remote unit transmits a reverse link signal which arrives with a power level equal to the power level received from every other remote unit therefore having a low signal to noise ratio. Also, a pilot channel transmits a known sequence of data. Without the pilot signal, the searching process must examine all possibilities of what data may have been transmitted.

FIG. 2 shows an exemplary embodiment of a prior art base station. The base station of FIG. 2 has one or more antennas **12** receiving CDMA reverse link remote unit signals **14**. Typically, an urban base station's coverage area is split into three sub-regions called sectors. With two antennas per sector, a typical base station has a total of six receive antennas. The received signals are down-converted to baseband by analog receiver **16** that quantizes the received signal I and Q channels and sends these digital values over signal lines **18** to channel element modem **20**. A typical base station comprises multiple channel element modems like channel element modem **20** (not shown in FIG. 2). Each channel element modem **20** supports a single user. In the preferred embodiment, channel element modem **20** comprises four demodulation elements **22** and eight searchers **26**. Microprocessor **34** controls the operation of demodulation elements **22** and searchers **26**. The user PN code in each demodulation element **22** and searcher **26** is set to that of the remote unit assigned to that channel element modem **20**. Microprocessor **34** steps searchers **26** through a set of offsets, called a search window, that is likely to contain multipath signal peaks suitable for assignment of demodulation elements **22**. For each offset, searcher **26** reports the energy it finds at that offset to microprocessor **34**. Demodulation elements **22** are then assigned by microprocessor **34** to the paths identified by searchers **26**. Once one of demodulation elements **22** has locked onto the signal at its assigned offset, it then tracks that path on its own without supervision from microprocessor **34**, until the path fades away or until it is assigned to a new path by microprocessor **34**.

For the system of FIG. 2, each demodulation element **22** and searcher **26** contains one FHT processor **52** capable of performing one FHT transform during a time period equal to the period of a Walsh symbol. The FHT processor is slaved to "real time" in the sense that every Walsh symbol interval one value is input and one symbol value is output from the FHT.

Therefore, to provide a rapid searching process, more than one searcher **26** must be used. Each of searchers **26** supplies back to microprocessor **34** the results of the search it performs. Microprocessor **34** tabulates these results for use in the assignment of demodulation elements **22** to the incoming signals.

In FIG. 2, the internal structure of only one demodulation element **22** is shown, but should be understood to apply to searchers **26** as well. Each demodulation element **22** or searcher **26** of the channel element modem has a corresponding I PN and Q PN sequence generator **36, 38** and the user-specific PN sequence generator **40** that is used to select a particular remote unit. User-specific PN sequence output **40** is XOR'd by XOR gates **42** and **44** with the output of I PN and Q PN sequence generators **36** and **38** to produce PN-I' and PN-Q' sequences that are provided to despreader **46**. The timing reference of PN generators **36, 38, 40** is adjusted to the offset of the assigned signal, so that despreader **46** correlates the received I and Q channel antenna samples with the PN-I' and PN-Q' sequence consistent with the assigned signal offset. Four of the despreader outputs, corresponding to the four PN chips per Walsh chip, are summed to form a single Walsh chip by accumulators **48** and **50**. The accumulated Walsh chip is then input into Fast Hadamard Transform (FHT) processor **52**. When 64 chips corresponding to one Walsh symbol have been received, FHT processor **52** correlates the set of sixty-four Walsh chips with each of the sixty-four possible transmitted Walsh symbols and outputs a sixty-four entry matrix of soft decision data. The output of FHT processor **52** is then combined with those of other assigned demodulation elements by combiner **28**. The output of combiner **28** is a "soft decision" demodulated symbol, weighted by the confidence that it correctly identifies the originally transmitted Walsh symbol. The soft decision data is then passed to forward error correction decoder **29** for further processing to recover the original call signal. This call signal is then sent through digital link **30**, such as a T1 or E1 link, that routes the call to public switched telephone network (PSTN) **32**.

Like each demodulation element **22**, each searcher **26** contains a demodulation data path with an FHT processor capable of performing one FHT transform during a time period equal to the period of a Walsh symbol. Searcher **26** only differs from demodulation element **22** in how its output is used and in that it does not provide time tracking. For each offset processed, each searcher **26** finds the correlation energy at that offset by despreading the antenna samples, accumulating them into Walsh chips that are input to the FHT transform, performing the FHT transform and summing the maximum FHT output energy for each of the Walsh symbols for which the searcher dwells at an offset. The final sum is reported back to microprocessor **34**.

Generally each searcher **26** is stepped through the search window with the others as a group by microprocessor **34**, each separated from its neighbor by half of a PN chip. In this way enough correlation energy exists at each maximum possible offset error of a quarter chip to ensure that a path is not missed because the searcher did not correlate with the exact offset of the path. After sequencing searchers **26** through the search window, microprocessor **34** evaluates the results reported back, looking for strong paths for demodulation elements assignment as described in above mentioned co-pending U.S. Patent Application Serial No. 08/144,902.

The multipath environment is constantly changing as the remote unit and other reflective objects move about in the base station coverage area. The number of searches that must be performed is set by the need to find multipath quickly enough so that valid paths may be put to good use by the demodulation elements. On the other hand, the number of demodulation elements required is a function of the number of paths generally found to be usable at any point in time. To meet these needs, the system of FIG. 2 has two searchers **26** and one demodulation element **22** for each of four demodulator integrated circuits (IC's) used, for a total of four demodulation elements and eight searchers per channel element modem. Each of these twelve processing elements contains a complete demodulation data path, including the FHT processor which takes a relatively large, costly amount of area to implement on an integrated circuit. In addition to the four demodulator IC's the channel element modem also has a modulator IC and a forward error correction decoder IC for a total of 6 IC chips. A powerful and expensive microprocessor is needed to manage and coordinate the demodulation elements and the searchers. As implemented in FIG. 2, these circuits are completely independent and require the close guidance of microprocessor **34** to sequence through the correct offsets, and handle the FHT outputs. Every Walsh symbol microprocessor **34** receives an interrupt to process the FHT outputs. This interrupt rate alone necessitates use of a high powered microprocessor.

It would be advantageous if the six IC's required for a modem could be reduced to a single IC needing less microprocessor support, thereby reducing the direct IC cost and board-level production cost of the modem, and allowing migration to a lower cost microprocessor (or alternately a single high powered microprocessor supporting several channel element modems at once.) Just relying on shrinking feature sizes of the IC fabrication process and placing the six chips together on a single die is not enough. The fundamental architecture of the searcher needs to be redesigned for a truly cost effective single chip modem. From the discussion above, it should be apparent that

there is a need for a signal receiving and processing apparatus that can demodulate a spread spectrum call signal at a lower cost and in a more architecturally efficient manner.

The present invention can use a set of real time searchers as described above or a single, integrated search processor that can quickly evaluate large numbers of offsets that potentially contain multipath of a received call signal.

The present invention is a method of searching for a multipath signal which is transmitted at an unknown variable rate and is subjected to power control.

SUMMARY OF THE INVENTION

The present invention is a method of searching for a multipath signal which is transmitted at an unknown variable rate and is subjected to power control. The searching method is linear in that no attempt is made to synchronize the searching process to time known to contain data. The searching process is aligned to power control group boundaries so that accurate power estimates can be made.

BRIEF DESCRIPTION OF THE DRAWINGS

The features, objects, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

FIG. 1 represents an exemplary severe multipath signal condition;

FIG. 2 is a block diagram of a prior art communications network demodulation system;

FIG. 3 represents an exemplary CDMA telecommunications system constructed in accordance with the present invention;

FIG. 4 is a block diagram of a channel element modem constructed in accordance with the present invention;

FIG. 5 is a block diagram of the search processor;

FIG. 6 illustrates the circular nature of the antenna sample buffer using a first offset;

FIG. 7 illustrates the circular nature of the antenna sample buffer for a second accumulation at the first offset of FIG. 6;

FIG. 8 illustrates the circular nature of the antenna sample buffer for a second offset;

FIG. 9 is a graph showing how the searcher processes the receiver input as a function of time;

FIG. 10 is a block diagram of the searcher front end;

FIG. 11 is a block diagram of the searcher despreader;

FIG. 12 is a block diagram of the searcher result processor;

FIG. 13 is a block diagram of the searcher sequencing control logic;

FIG. 14 is a timing diagram showing the processing sequence depicted in FIG. 5, showing the corresponding states of certain control logic elements presented in FIG. 13; and

FIG. 15 is an alternative block diagram of the search processor.

DESCRIPTION OF THE PREFERRED EMBODIMENT

In the following description of a method and system for processing telephone calls within a digital wireless telephone system, various references are made to processes and steps that are performed in order to achieve a desired result. It should be understood that such references do not describe human actions or thought, but are directed towards the operation, modification and transformation of various systems including and especially those systems which process electrical and electromagnetic signals and charges, optical signals, or a combination thereof. Fundamental to such systems is the use of various information storage devices, often referred to as "memory", which store information via the placement and organization of atomic or super-atomic charged particles on hard disk media or within silicon, gallium arsenic, or other semiconductor based integrated circuit media, and the use of various information processing devices, often referred to as "microprocessors," which alter their condition and state in response to such electrical and electromagnetic signals and charges. Memory and microprocessors that store and process light energy or particles having special optical characteristic, or a combination thereof, are also contemplated and use thereof is consistent with the operation of the described invention.

The present invention can be implemented in a wide variety of data transmission applications and in the preferred embodiment illustrated in FIG. 3 is implemented within system **100** for voice and data transmission in which system controller and switch (BSC&S) **102** performs interface and control functions to permit call communications with remote units **104** through base stations **106**. BSC&S **102** controls the routing of calls between public switched telephone network (PSTN) **108** and base stations **106** for transmission to and from remote units **104**.

FIG. 4 illustrates channel element modems **110A - 110N** and other elements of the base station infrastructure operating in accordance with the CDMA methods and data formats described in the above-referenced patents.

A plurality of antennas **112** provides received reverse link signal **114** to analog transmitter receiver (transceiver) **116**. Analog transceiver **116** down-converts reverse link signal **114** to baseband and samples the baseband waveform at eight times the PN chip rate of the CDMA received signal as defined above. Analog transceiver **116** provides the digital antenna samples to channel element modems **110A - 110N** through base station RX backplane signal **118**. Each channel element modem **110A - 110N** may be assigned to one remote unit having an active communication established with the base station. Each channel element modem **110A - 110N** is nearly identical in structure.

When channel element modem **110A** is assigned to an active call, demodulator front end **122** and integrated search processor **128** isolate a signal from the corresponding remote unit from the plurality of call signals contained in reverse link signal **114** by use of the PN sequences as described in the above referenced patents and patent applications. Channel element modem **110A** includes single integrated search processor **128** to identify multipath signals that can be used by demodulator front end **122**. In the preferred embodiment, time sliced FHT processor engine **120** services both integrated search processor **128** and demodulator front end **122**. Other than sharing FHT processor engine **120** and related max detect block **160**, integrated search processor **128**, is stand-alone, self-controlled, and self-contained. A searching architecture is detailed in a co-pending U.S. Patent Application Serial No. 08/316,177 entitled "MULTIPATH SEARCH PROCESSOR FOR A SPREAD SPECTRUM MULTIPLE ACCESS COMMUNICATION SYSTEM" filed September 30, 1994, and assigned to the assignee of the present invention.

FHT processor engine **120** is the core of the demodulation process. In the preferred embodiment, FHT processor engine **120** correlates the received Walsh symbol values with each of the possible Walsh symbols that may have been transmitted by the remote unit. FHT processor engine **120** outputs a correlation energy corresponding to each possible Walsh symbol where a higher correlation energy level corresponds to a higher likelihood that the symbol corresponding to that Walsh index was communicated by the remote unit. Max detect block **160** then determines the largest of the 64 FHT transform energy outputs. The maximum correlation energy and the corresponding Walsh index from max detect block **160** and each of the 64 correlation energy output from FHT processor engine **120** are passed to pipelined demodulator processor **126** for further signal processing. The maximum correlation energy and the corresponding Walsh index from max detect block **160** are passed back to integrated search processor **128**.

Pipelined demodulator processor **126** time aligns and combines symbol data received at different offsets into a single demodulated "soft decision" symbol stream. In addition, pipelined demodulator processor **126** calculates the power level of the signal being received. From the received power level a power control indication is created to command the remote unit to raise or lower the remote unit's transmit power. The power control indication is passed through modulator **140** which adds the indication to the base station transmitted signal for reception by the remote unit. This power control loop operates under the method described in U.S. Patent 5,056,109 referenced above.

The soft decision symbol stream from pipelined demodulator processor **126** is output to deinterleaver/forward error correction decoder **130** where it is deinterleaved and decoded. Channel element microprocessor **136** supervises the entire demodulation procedure and obtains the recovered data from deinterleaver/forward error correction decoder **130** via microprocessor bus interface **134**. The data is then routed through digital backhaul link **121** to BSC&S **102** which connects the call through PSTN **108**.

The forward link data path proceeds much as the inverse of the functions just presented for the reverse link. The signal is provided from PSTN **108** through BSC&S **102** and to digital backhaul **121**. Digital backhaul **121** provides input to encoder/interleaver **138** through channel element microprocessor **136**. After encoding and interleaving the data, encoder/interleaver **138** passes the data to modulator **140** where it is modulated as disclosed in the above referenced patents. Output **146** of modulator **140** is passed to transmit summer **142** where it is added to the outputs of other channel element modems **110B - 110N** prior to being up converted from baseband and amplified in analog transmitter receiver **116**. A summing method is disclosed in a co-pending U.S. Patent Application Serial No. 08/316,156 entitled "SERIAL LINKED INTERCONNECT FOR THE SUMMATION OF MULTIPLE DIGITAL WAVEFORMS," filed September 30, 1994, and assigned to the assignee of the present invention. As presented in the above referenced patent application, the transmit summer corresponding to each of channel element modems **110A - 110N** can be cascaded in a daisy-chain fashion eventually resulting in a final sum that is provided to analog transceiver **116** for broadcasting.

FIG. 5 shows the elements comprising integrated search processor **128**. The heart of the searching process is time sliced FHT processor engine **120**, which, as mentioned above, is shared between integrated search processor **128** and demod front end **122** (not shown in FIG. 5). FHT processor engine **120** can

perform Walsh symbol transforms at a rate 32 times faster than FHT processor **52** of FIG. 2. This rapid transform capability empowers the time sliced operation of channel element modem **110**.

In the preferred embodiment FHT processor engine **120** is constructed using a six stage butterfly network. As explained in detail above, a Walsh function of order n can be defined recursively as follows:

$$W(n) = \begin{vmatrix} W(n/2)\ ,\ W(n/2) \\ W(n/2)\ ,\ W'(n/2) \end{vmatrix}$$

where W' denotes the logical complement of W, and W(1) = 0.

In the preferred embodiment a Walsh sequence is generated where n = 6, therefore a 6-stage butterfly trellis is used to correlate the 64 Walsh chips of one transmitted Walsh symbol with each of the 64 possible Walsh sequences. A structure and method of operation for FHT processor engine **120** is detailed in a co-pending U.S. Patent Application Serial No. 08/173,460 entitled "METHOD AND APPARATUS FOR PERFORMING A FAST HADAMARD TRANSFORM," filed December 22, 1993, assigned to the assignee of the present invention.

To reap the benefits of FHT processor engine **120** with thirty-two times the throughput of its real-time-slaved counterpart, FHT processor engine **120** must be provided with high rate input data to process. Antenna sample buffer **172** has been specifically tailored to meet this need. Antenna sample buffer **172** is written to and read from in a circular manner

The searching process is grouped in sets of single offset searches. The highest level of grouping is the antenna search set. Each antenna search set is made up of a plurality of search windows. Typically each search window in the antenna search set is an identically performed search group where each search window in the antenna search receives data from a different antenna. Each search window is made up of a series of search rakes. A search rake is a set of sequential search offsets that is performed in a time equivalent to the duration of a Walsh symbol. Each search rake is comprised of a set of rake elements. Each rake element represents a single search at a given offset.

At the beginning of the searching process, channel element microprocessor **136** sends parameters specifying a search window which may be part of an antenna search set. The width of the search window may be designated in PN chips. The number of search rakes needed to complete the search window varies depending on the number of PN chips specified in the

search window. The number of rake elements per search rake can be specified by channel element microprocessor **136** or could be fixed to some constant.

Referring again to FIG. 1 showing an exemplary set of signals arriving at the base station from a single remote unit, the relationship of the search window, search rake, and rake element becomes more clear. The vertical axis in FIG. 1 represents the power received in decibels (dB). The horizontal axis represents the delay in the arrival time of a signal due to multipath delays. The axis (not shown) going into the page represents a segment of time. Each signal spike in the common plane of the page has arrived at the same time but has been transmitted by the remote unit at a different time.

The horizontal axis can be thought of as having units of PN chip offset. At any given time, the base station perceives a variety of signals from a single remote unit, each of which has traveled a different path and may have a different delay than the others. The remote unit's signal is modulated by a PN sequence. A copy of the PN sequence is also generated at the base station. At the base station, if each multipath signal were individually demodulated, a PN sequence code aligned to each signal's timing would be needed. Each of these aligned PN sequences would be delayed from the zero offset reference at the base station due to the delay. The number of PN chips that the aligned PN sequence is delayed from the zero offset base station reference could be mapped to the horizontal axis.

In FIG. 1, time segment **10** represents a search window set of PN chip offsets to be processed. Time segment **10** is divided into five different search rakes such as search rake time segment **9**. Each search rake is in turn made up of a number of rake elements which represent the actual offsets to be searched. For example, in FIG. 1, each search rake is made up of 8 different rake elements such as the rake element indicated by arrow **8**.

To process a single rake element as indicated by arrow **8**, a set of samples over time at that offset are needed. For example, to process the rake element indicated by arrow **8**, the despreading process needs the set of sample at the offset indicated by arrow **8** going back into the page over time. The despreading process also needs a corresponding PN sequence. The PN sequence can be determined by noting the time the samples arrived and the offset desired to be processed. The desired offset can be combined with the arrival time to determine the corresponding PN sequence to be correlated with the received samples.

As the rake element is despread the receive antenna samples and the PN sequence are run through a series of values over time. Note that the received antenna samples are the same for all offsets shown in FIG. 1 and

spikes 2-7 are showing exemplary multipath peaks which arrive simultaneously and are only discriminated by the despreading process.

In the preferred embodiment described below, each rake element is offset in time from the preceding rake element by one half PN chip in time. This means that if the rake element corresponding to arrow 8 was correlated beginning from the sliced plane shown and moving forward in time (into the page as shown) then the rake element to the left of the one corresponding to arrow 8 would use samples starting one half chip in time back from the sliced plane shown. This progression in time allows each rake element in a common search rake to be correlated to the same PN sequence.

Each remote unit receives the base station's transmitted signal delayed by some amount due to the path delay through the terrestrial environment. The same I and Q PN short code and the user PN long code generation is also being performed in the remote unit. The remote unit generates a time reference based on the time reference it perceives from the base station. The remote unit uses the time reference signal as an input to its I and Q PN short code and the user PN long code generators. The information signal received at the base station from the remote unit is therefore delayed by the round trip delay of the signal path between the base station and the remote unit. Therefore if the timing of the PN generator used in the searching process is slaved to the zero offset timing reference at the base station, the output of the generators is always available before the corresponding signal is received from the remote unit.

In an OQPSK signal, the I channel data and the Q channel data are offset from each other by one half chip in time. Therefore OQPSK despreading used in the preferred embodiment requires data sampled at twice the chip rate. The searching process also operates optimally with data sampled at half the chip rate. Each rake element within a search rake is offset by one half chip from the previous rake element. The one half chip rake element resolution ensures that multipath peak signals are not skipped over without detection. For these reasons antenna sample buffer 172 of FIG. 5 stores data sampled at twice the PN chip rate.

One Walsh symbol worth of data is read from antenna sample buffer 172 to process a single rake element. For each successive rake element, one Walsh symbol worth of data is read out of antenna sample buffer 172 one half of a PN chip offset from the previous rake element. Each rake element is despread with the same PN sequence read from PN sequence buffer 176 by despreader 178 for each rake element in the search rake.

Antenna sample buffer **172** is two Walsh symbols deep and is continually and repeatedly read from and written to throughout the searching process. Within each search rake, the rake element having the latest offset in time is processed first. The latest offset corresponds to the signal which has traveled the longest signal path from the remote unit to the base station. The time at which the searcher starts to process a search rake is keyed to the Walsh symbol boundaries associated with the rake element having the latest offset in the search rake. A time strobe, referred to as the offset Walsh symbol boundary, indicates the earliest time that all of the samples needed are available in antenna sample buffer **172** and the searching process can begin the first rake element in the search rake.

The operation of antenna sample buffer **172** is most easily illustrated by noting its circular nature. FIG. 6 shows an illustrative diagram of the operation of antenna sample buffer **172**. In FIG. 6 thick circle **400** can be thought of as antenna sample buffer **172** itself. Antenna sample buffer **172** contains memory locations for two Walsh symbols worth of data. Write pointer **406** circulates around antenna sample buffer **172** in the direction indicated in real time, meaning that write pointer **406** rotates around the two Walsh symbol deep antenna sample buffer **172** in the time that two Walsh symbols worth of samples are passed to searcher front end **174**. As the samples are written into antenna sample buffer **172** according to the memory location indicated by write pointer **406**, the previously stored values are overwritten. In the preferred embodiment, antenna sample buffer **172** contains 1024 antenna samples because each of the two Walsh symbols contains 64 Walsh chips, each Walsh chip contains 4 PN chips, and each PN chip is sampled twice.

The operation of the searching process is divided into discrete 'time slices.' In the preferred embodiment, a time slice is equal to 1/32 of the Walsh symbol duration. The choice of 32 time slices per Walsh symbol is derived from the available clocking frequency and number of clock cycles need to perform an FHT. 64 clock cycles are required to perform an FHT for one Walsh symbol. In the preferred embodiment, a clock running at eight times the PN chip frequency is available and provides the necessary performance level. Eight times the PN chip rate multiplied by the 64 required clocks is equivalent to the time it takes to receive two Walsh chips worth of data. Because there are 64 Walsh chips in each half of the buffer, 32 time slices are needed to read in a complete Walsh symbol.

In FIG. 6, a set of concentric arcs outside of thick circle **400** represents read and write operation with antenna sample buffer 172. (The arcs within

thick circle **400** are used to aid explanation and do not correspond to read or write operations.) Each arc represents a read or write operation during one time slice. The arc closest to the center of the circle occurs first in time and each successive arc represents an operation occurring in successively later time slices as indicated by time arrow **414**. Each of the concentric arcs corresponds to a section of antenna sample buffer **172** as represented by thick circle **400**. If one were to imagine radii drawn from the center of thick circle **400** to the end points of each of the concentric arcs, the portion of thick circle **400** between the intersection of the radii and thick circle **400** would be representative of the memory locations accessed. For example, during the first time slice operation shown, 16 antenna samples are written to antenna sample buffer **172** represented by arc **402A**.

In FIGS 6, 7, and 8 the following search parameters for the illustrative search window are assumed:

Search window width = 24 PN chips
Search offset = 24 PN chips
Number of symbols to accumulate = 2
Number of rake elements per search rake =24.

FIG. 6 also assumes that antenna sample buffer **172** contains nearly a full Walsh symbol worth of valid data before the write indicated by arc **402A**. During subsequent time slices, a write corresponding to arc **402B** and to arc **402C** occurs. During the 32 time slices available during one Walsh symbol worth of time, the write operations continue from arc **402A** to arc **402FF** most of which are not shown.

The 32 time slices represented by arcs **402A** to **402FF** correspond to the time used to complete one search rake. Using the parameters given above, the search rake begins 24 PN chips offset from zero offset reference or 'real time' and contains 24 rake elements. The 24 PN chip offset corresponds to a rotation 16.875 degrees around thick circle **400** from the beginning of the first write indicated by arc **402A** (calculated by dividing the 24 PN chip offset by the 256 total number of chips in half antenna sample buffer **172** and multiplying by 180 degrees.) The 16.875 degree arc is illustrated by arc **412**. The 24 rake elements correspond to reads indicated by arcs **404A - 404X** most of which are not shown. The first read corresponding to arc **404A** begins at the search offset some time after the write corresponding to **402C** so that a contiguous set of data is available. Each successive read such as **404B** is offset from the previous by a single memory location, corresponding to a 1/2 PN chip of time. During the search rake shown, the reads move toward earlier time offsets as shown by arcs **404A - 404X** slanting counter clockwise with

progressing time in the opposite rotation direction as write pointer indication **406**. The 24 read represented by arcs **404A** to **404X** traverse the arc indicated by arc **418**. The progression of the reads toward earlier samples has the advantage of providing seamless searching within a search window as each search rake is executed. This advantage is explained in detail subsequently herein.

Each of the reads corresponding to arcs **404A** to **404X** passes one Walsh symbol worth of data to despreader **178**. The read therefore corresponds to traversing thick circle **400** by 180 degrees. Note that in the search rake shown in FIG. 6, the last write corresponding to arc **402FF**, and last read corresponding to arc **404X** do not include any common memory locations to ensure contiguous valid data. However, hypothetically, if the pattern of read and writes were to continue they would in fact intersect and valid data would not be provided under this condition.

In most signaling conditions, the result of a rake element worth of data collected during one Walsh symbol worth of time is not sufficient to provide accurate information about the location of diverse signals. In these cases, a search rake may be repeated multiple times. The results of rake elements in successive search rakes at a common offset are accumulated by search result processor **162** as explained in detail subsequently herein. In this case the search parameters given above indicate that the number of symbols to accumulate at each offset is two. FIG. 7 shows the search rake of FIG. 6 repeated at the same offset for the next successive Walsh symbol worth of data. Note that antenna sample buffer **172** contains two Walsh symbols worth of data so that the data that is needed for processing during the search rake indicated on FIG. 7 was written during the search rake shown on FIG. 6. In this configuration, memory locations 180 degrees away from each other represent the same PN offset.

After completing the two accumulated search rakes in FIGS 6 and 7, the searching process advances to the next offset in the search window. The amount of the advance is equal to the width of the search rake processed, in this case 12 PN chips. As specified in the search parameters, the search window width is 24 PN chips. The width of the window will determine how many search rake offsets are needed to complete the search window. In this case two different offsets are needed to cover the 24 PN chip window width. The window width is indicated on FIG. 8 by arc **412**. The second offset for this search window begins at the offset following the last offset of the previous search rake and continues around to the nominal zero offset point as set by the location of the beginning of the first write as indicated by arc **430A**. Again

there are 24 rake elements within the search rake as indicated by arcs **432A - 432X** most of which are not shown. Again the 32 writes are indicated by the arcs **430A - 430FF**. Thus the last write, as indicated by arc **430FF**, and the last read, as indicated by arc **432X**, abut one another in antenna sample buffer **172**.

The search rake shown in FIG. 8 is repeated on the opposite side of antenna sample buffer **172** much as the search rake in FIG. 6 is repeated in FIG. 7 because the search parameters designate that each symbol is accumulated twice. At the completion of the second accumulation of the second search rake, integrated search processor **128** is available to begin another search window. The subsequent search window could have a new offset or it could specify a new antenna or both.

In FIG. 8, the location of the boundary between the read half and the write half of the buffer is marked with label **436**. In FIG. 6, the boundary is marked with label **410**. The signal which indicates the point in time corresponding to labels **410** and **436** is referred to as the offset Walsh symbol strobe and also indicates that a new Walsh symbol worth of samples is available. As the search rakes within a window advance to earlier offsets, the boundary between the read and write halves of the buffer slews in lock step counterclockwise as shown in FIG. 8. If after the completion of the present search window, if a large change in the offset being processed is desired, the offset Walsh symbol strobe may be advanced a large portion of the circumference of the circle.

FIG. 9 is a search timeline that provides further graphical illustration of the searcher processing. Time is plotted along the horizontal axis in units of Walsh symbols. Antenna sample buffer **172** addresses and PN sequence buffer **176** addresses are shown along the vertical axis, also in units of Walsh symbols. Because antenna sample buffer **172** is two Walsh symbols deep, antenna sample buffer **172** addressing wraps on even Walsh symbol boundaries, but for illustrative purposes, FIG. 9 shows the addresses before being folded on top of one another. Samples are written into antenna sample buffer **172** at an address taken directly from the time they were obtained, so write pointer **181** into antenna sample buffer **172** is a straight forty-five degree inclined line. The offset being processed maps into a base address in antenna sample buffer **174** to start a read of a Walsh symbol of samples for a single rake element. The rake elements are illustrated in FIG. 9 as nearly vertical read pointer line segments **192**. Each rake element maps to a Walsh symbol in height as referred to the vertical axis and 1/32 of a Walsh symbol as referred to the horizontal axis.

The vertical gaps between the rake elements within a search rake are caused by demod front end **122** interrupting the search process to use FHT processor engine **120**. Demod front end **122** operates in real time and has first priority use of FHT processor engine **120** whenever it has a current or queued set of data for processing. Therefore typically use of FHT processor engine **120** is given to demod front end **120** on each Walsh symbol boundary corresponding to a PN offset that is being demodulated by demod front end **122**.

FIG. 9 shows the same search rakes shown in FIGS. 6, 7, and 8. For example, search rake **194** has 24 rake elements each of which corresponds to one to the read arcs **404A** - **404X** on FIG. 6. On FIG. 9 for search rake **194**, pointer **410** indicates the offset Walsh symbol strobe corresponds to the like pointer on FIG. 6. To read the current samples, each rake element must be beneath write pointer **181**. The downward slope of the rake elements with a search rake indicates the steps towards earlier samples. Search rake **195** corresponds to the search rake shown in FIG. 7 and search rake **196** corresponds to the search rake shown in FIG. 8.

In the search window defined by the parameters above, only 24 rake elements per search rake are specified even though the search rake has 32 available time slices. Each rake element can be processed in one time slice. However, it is not practically possible to increase the number of rake elements per search rake to 32 to match the number of time slices available during a search rake. Demod front end **122** uses some of the available time slices of the FHT processor. There is also a time delay associated with a rake advance as the read process must wait for the write process to fill the buffer with valid data at the advanced offset. Also some margin is needed to synchronize to a time slice processing boundary after observing the offset Walsh symbol strobe. All these factors practically limit the number of rake elements which can be processed in a single search rake. In some cases the number of rake elements per search rake could be increased such as if demod front end **122** has only one demodulation element assigned and hence only interrupts FHT processor engine **120** once per search rake. Therefore in the preferred embodiment, the number of rake elements per search rake is controllable by channel element microprocessor **136**. In alternative embodiments, the number of rake elements per search rake could be a fixed constant.

There also can be significant overhead delay when switching between source antennas at the input to the sample buffer or changing the search window starting point or width between searches. If one rake needs a particular set of samples and the next rake for a different antenna needs to use

an overlapping part of the buffer, the next rake must postpone processing until the next offset Walsh symbol boundary occurs, at which point a complete Walsh symbol of samples for the new antenna source is available. In FIG. 9, search rake **198** is processing data from a different antenna than search rake **197**. Horizontal line **188** indicates the memory location corresponding to the new antenna input samples. Note that search rake **197** and **198** do not use any common memory locations.

For every time slice, two Walsh chips of samples must be written to the sample buffer and one full Walsh symbol of samples may be read from the sample buffer. In the preferred embodiment, there are 64 clock cycles during each time slice. A full Walsh chip of samples is comprised of four sets of samples: ontime I channel samples, late I channel samples, ontime Q channel samples and late Q channel samples. In the preferred embodiment, each sample is four bits. Therefore sixty four bits per clock are needed from antenna sample buffer **172**. Using a single port RAM, the most straightforward buffer design doubles the word width to 128 bits, and splits the buffer into two 64 bit wide, 64 word, independently read/writeable even and odd Walsh chip buffers **168, 170**. The much less frequent writes to the buffer are then multiplexed in between reads, which toggle between the two banks on successive clock cycles.

The Walsh chip samples read from the even and odd Walsh chip buffers **168, 170** has an arbitrary alignment to the physical RAM word alignment. Therefore on the first read of a time slice, both halves are read into despreader **178** to form a two Walsh chip wide window from which the single Walsh chip with the current offset alignment is obtained. For even Walsh chip search offsets, the even and odd Walsh chip buffer address for the first read are the same. For odd Walsh chip offsets, the even address for the first read is advanced by one from the odd address to provide a consecutive Walsh chip starting from the odd half of the sample buffer. The additional Walsh chips needed by despreader **178** can be passed thereto by a read from a single Walsh chip buffer. Successive reads then ensure that there is always a refreshed two Walsh chip wide window from which to draw a Walsh chip of data aligned to the current offset being processed.

Referring again to FIG. 5, for each rake element in a search rake, the same Walsh symbol of PN sequence data from PN sequence buffer **176** is used in the despreading process. For every clock cycle of a time slice, four pairs of PN-I' and PN-Q' are needed. Using a single port RAM, the word width is doubled and read from half as often. The single write to PN sequence

buffer **176** needed per time slice is then performed on a cycle not used for reading.

Because the searching process can specify searching PN offsets of up to two Walsh symbols delayed from the current time, four Walsh symbols worth of PN sequence data must be stored. In the preferred embodiment PN sequence buffer **176** is a one hundred and twenty eight word by sixteen bit RAM. Four Walsh symbols are required because the starting offset can vary by 2 Walsh symbols and once the starting offset is chosen, one Walsh symbol worth of PN sequence is need for correlation meaning three Walsh symbols worth of data is need for the despreading process. Because the same PN sequence is repeatedly used, the data in PN sequence buffer **176** cannot be overwritten during the despreading process corresponding to a single search rake. Therefore an additional Walsh symbol worth of memory is needed to store the PN sequence data as it is generated.

The data that is written into both PN sequence buffer **176** and antenna sample buffer **172** is provided by searcher front end **174**. A block diagram of searcher front end **174** is shown in FIG. 10. Searcher front end **174** includes short code I and Q PN generators **202, 206** and the long code User PN generator **204**. The values output by short code I and Q PN generators **202, 206** and the long code User PN generator **204** are determined in part by the time of day. Each base station has a universal timing standard such as GPS timing to create a timing signal. Each base station also transmits its timing signal over the air to the remote units. At the base station, the timing reference is said to have zero offset because it is aligned to the universal reference.

The output of long code User PN generator **204** is logically XOR'd with the output of short code I and Q PN generators **202, 206** by XOR gates **208** and **210** respectively. (This same process is also performed in the remote unit and the output is used to modulate the remote unit's transmitted signal.) The output of XOR gates **208** and **210** is stored in serial to parallel shift register **212**. Serial to parallel shift register **212** buffers the sequences up to the width of PN sequence buffer **176**. The output of serial to parallel shift register **212** is then written into PN sequence buffer **176** at an address taken from the zero offset reference time. In this way, searcher front end **174** provides the PN sequence data to PN sequence buffer **176**.

Searcher front end **174** also provides antenna samples to antenna sample buffer **172**. Receive samples **118** are selected from one of a plurality of antennas via a MUX **216**. The selected receive samples from MUX **216** are passed to latch **218** where they are decimated, meaning one quarter of the samples are selected for use in the searching process. Receive samples **118**

have been sampled at eight times the PN chip rate by analog transmitter receiver **116** (of FIG. 4). Processing within the searching algorithm is designed for samples taken at one half the chip rate. Therefore only one quarter of the received samples need be passed to antenna sample buffer **172**.

The output of the latch **218** is fed to serial to parallel shift register **214**, which buffers the samples up to the width of antenna sample buffer **172**. The samples are then written into even and odd Walsh chip buffers **168, 170** at addresses also taken from the zero offset reference time. In this way, despreader **178** can align the antenna sample data with a known offset with respect to the PN sequence.

Referring again to FIG. 5, for each clock cycle in a time slice, despreader **178** takes a Walsh chip of antenna samples from antenna sample buffer **172** and a corresponding set of PN sequence values from PN sequence buffer **176** and outputs an I and Q channel Walsh chip to FHT processor engine **120** through MUX **124**.

FIG. 11 shows a detailed block diagram of despreader **178**. Even Walsh chip latch **220** and odd Walsh chip latch **222** latch the data from even Walsh chip buffer **168** and odd Walsh chip buffer **170** respectively. MUX bank **224** extracts the Walsh chip of samples to be used from the two Walsh chips worth of samples presented by even and odd Walsh chip latches **220** and **222**. MUX select logic **226** defines the boundary of the selected Walsh chip based on the offset of the rake element being processed. A Walsh chip is output to OQPSK despreader XOR bank **228**.

The PN sequence values from PN sequence buffer **176** are latched by PN sequence latch **234**. Barrel shifter **232** rotates the output of PN sequence latch **234** based on the offset of the rake element being processed and passes the PN sequence to OQPSK despreader XOR bank **228** which conditionally inverts the antenna samples based on the PN sequence. The XOR'd values are then summed through adder tree **230** which performs the sum operation in the OQPSK despread, and then sums four despread chip outputs together to form a Walsh chip for input to FHT processor engine **120**.

Referring again to FIG. 5, FHT processor engine **120** takes sixty-four received Walsh chips from despreader **178** through MUX **124**, and using a 6-stage butterfly trellis, correlates these sixty-four input samples with each of the sixty-four Walsh functions in a sixty-four clock cycle time slice. Max detect **160** can be used to find the largest of the correlation energies output from FHT processor engine **120**. The output of MAX detect **160** is passed on to search result processor **162** which is part of integrated search processor **128**.

Search result processor **162** is detailed in FIG. 12. Search result processor **162** also operates in a time sliced manner. The control signals provided to it are pipeline delayed to match the two time slice delay from the start of inputting Walsh chips to FHT processor engine **120** to obtaining the maximum energy output. As explained above, a set of search window parameters may designate that a number of Walsh symbols worth of data be accumulated before the results of the chosen offset are processed. In the parameters used with the example of FIGS. 6, 7, 8, and 9, the number of symbols to accumulate is 2. Search result processor **162** performs the summing function along with other functions.

As search result processor **162** performs the sums over consecutive Walsh symbols, it must store a cumulative sum for each rake element in the search rake. These cumulative sums are stored in Walsh symbol accumulation RAM **240**. The results of each search rake are input to summer **242** from max detect **160** for each rake element. Summer **242** sums the present result with the corresponding intermediate value available from Walsh symbol accumulation RAM **240**. On the final Walsh symbol accumulation for each rake element, the intermediate result is read from Walsh symbol accumulation RAM **240** and summed by summer **242** with the final energy from that rake element to produce a final search result for that rake element offset. The search results are then compared with the best results found in the search up to this point as explained below.

In the above mentioned co-pending U.S. Patent Application Serial No. 08/144,902 entitled "DEMODULATION ELEMENT ASSIGNMENT IN A SYSTEM CAPABLE OF RECEIVING MULTIPLE SIGNALS," the preferred embodiment assigns the demodulation elements based on the best results from a search. In the present preferred embodiment, the eight best results are stored in best result register **250**. (A lesser or greater number of results could be stored in other embodiments.) Intermediate result register **164** stores the peak values and their corresponding rank order. If the current search result energy exceeds at least one of the energy values in intermediate result register **164**, search result processor control logic **254** discards the eighth best result in intermediate result register **164**, and inserts the new result, along with its appropriate rank, the PN offset, and antenna corresponding to the rake element result. All lesser ranked results are "demoted" one ranking. There are a great number of methods well known in the art for providing such a sorting function. Any one of them could be used within the scope of this invention.

Search result processor **162** has a local peak filter basically comprised of comparator **244** and previous energy latch **246**. The local peak filter, if enabled, prevents intermediate result register **164** from being updated even though a search result energy would otherwise qualify for inclusion, unless the search result represents a local multipath peak. In this way, the local peak filter prevents strong, broad "smeared" multipath from filling multiple entries in intermediate result register **164**, leaving no room for weaker but distinct multipath that may make better candidates for demodulation.

The implementation of the local peak filter is straightforward. The energy value of the previous rake element summation is stored in previous energy latch **246**. The present rake element summation is compared to the stored value by comparator **244**. The output of comparator **244** indicates which of its two inputs is larger and is latched in search result processor control logic **254**. If the previous sample represented a local maxima, search result processor control logic **254** compares the previous energy result with the data stored in intermediate result register **164** as described above. If the local peak filter is disabled by channel element microprocessor **136** then the comparison with intermediate result register **164** is always enabled. If either the leading or the last rake element at the search window boundary has a slope, then the slope latch is set so the boundary edge value can be considered as a peak as well.

The simple implementation of this local peak filter is aided by the progression of the reads toward earlier symbols within a search rake. As illustrated in FIGS. 6, 7, 8, and 9, within a search rake each rake element progress toward signals arriving earlier in time. This progression means that within a search window, the last rake element of a search rake and the first rake element of the subsequent search rake are contiguous in offset. Therefore, the local peak filter operation does not have to change and the output of comparator **244** is valid across search rake boundaries.

At the end of processing a search window, the values stored in intermediate result register **164** are transferred to best result register **250** readable by channel element microprocessor **136**. Search result processor **162** has thus taken much of the workload from channel element microprocessor **136**, which in the system of FIG. 2 needed to handle each rake element result independently.

The preceding sections have focused on the processing data path of integrated search processor **128** and have detailed how raw antenna samples **118** are translated into a summary multipath report at the output of

best result register **250**. The following sections detail how the each of the elements in the search processing data path are controlled.

Search control block **166** of FIG. 5 is detailed in FIG. 13. As mentioned previously, channel element microprocessor **136** specifies a search parameter set including the group of antennas to search over as stored in antenna select buffer **348**, the starting offset as stored in search offset buffer **308**, the number of rake elements per search rake as stored in rake width buffer **312**, the width of the search window as stored in search width buffer **314**, the number of Walsh symbols to accumulate as stored in Walsh symbol accumulation buffer **316**, and a control word as stored in control word buffer **346**.

The starting offset stored in search offset buffer **308** is specified with eighth chip resolution. The starting offset controls which samples are removed by decimation by latch **218** of FIG. 10 in searcher front end **174**. Due to the two Walsh symbol wide antenna sample buffer **172** in this embodiment, the largest value of the starting offset is half of a PN chip less than two full Walsh symbols.

Up until this point, the generic configuration to perform a search has been disclosed. In reality there are several classes of predefined searches. When a remote unit initially attempts to access the system, it sends a beacon signal called a preamble using the Walsh zero symbol. Walsh zero symbol is the Walsh symbol which contains all logical zeros instead of half ones and zeroes as described above. When a preamble search is performed, the searcher looks for any remote unit sending a Walsh zero symbol beacon signal on an access channel. The search result for a preamble search is the energy for the Walsh zero symbol. When an acquisition mode access channel search is performed, max detect **160** outputs the energy for Walsh zero symbol regardless of the maximum output energy detected. The control word stored in control word buffer **346** includes a preamble bit which indicates when a preamble search is being performed.

As discussed above, the power control mechanism of the preferred embodiment measures the signal level received from each remote unit and creates a power control indication to command the remote unit to raise or lower the remote unit's transmit power. The power control mechanism operates over a set of Walsh symbols called a power control group during traffic channel operation. (Traffic channel operation follows access channel operation and implies operation during an active call.) All the Walsh symbols within a single power control group are transmitted using the same power control indication command at the remote unit.

Also as described above, in the preferred embodiment of the present invention, the signal transmitted by the remote unit is of a variable rate during traffic channel operation. The rate used by the remote unit to transmit the data is unknown at the base station during the searching process. As the consecutive symbols are accumulated, it is imperative that the transmitter is not gated off during the accumulation. Consecutive Walsh symbols in a power control group are gated as a group meaning that the 6 Walsh symbols comprising a power control group in the preferred embodiment are all gated on or all gated off.

Thus when the search parameter specifies that a plurality of Walsh symbols be accumulated during traffic channel operation, the searching process must align each search rake to begin and end within a single power control group. The control word stored in control word buffer **346** includes a power control group alignment bit. With the power control group alignment bit set to one indicating a traffic channel search, the searching process synchronizes to the next power control group boundary instead of just the next offset Walsh symbol boundary.

The control word stored in control word buffer **346** also includes the peak detection filter enable bit as discussed earlier in conjunction with FIG. 8.

The searcher operates either in continuous or single step mode, according to the setting of the continuous/single step bit of the control word. In single step mode, after a search is performed, integrated search processor **128** returns to an idle state to await further instructions. In continuous mode, integrated search processor **128** is always searching, and by the time channel element microprocessor **136** is signaled that the results are available, integrated search processor **128** has started the next search.

Search control block **166** produces the timing signals used to control the searching process performed by integrated search processor **128**. Search control block **166** sends the zero offset timing reference to short code I and Q PN generators **202, 206** and long code User PN generator **204**, and the enable signal to decimator latch **218** and the select signal to MUX **216** in searcher front end **174**. It provides the read and write addresses for PN sequence buffer **176** and even and odd Walsh chip buffers **168** and **170**. It outputs the current offset to control the operation of despreader **178**. It provides the intra-time slice timing reference for FHT processor engine **120**, and determines whether the searching process or the demodulation process uses FHT processor engine **120** by controlling FHT input MUX **124**. It provides several pipeline delayed versions of certain internal timing strobes to search result processor control logic **254** of FIG. 12 to allow it to sum search results

across a rake of offsets for a number of Walsh symbol accumulations. Search control block **166** provides best result register **250** with the pipelined offset and antenna information corresponding to accumulated energy values stored.

In FIG. 13, system time count **342** is slaved to the zero offset time reference. In the preferred embodiment as previously detailed, the system clock runs at eight times the PN chip rate. There are **256** PN chips in a Walsh symbol, and 6 Walsh symbols in a power control group for a total of $6 \times 256 \times 8 = 12{,}288$ system clocks per power control group. Therefore in the preferred embodiment, system time count **342** is comprised of a fourteen bit counter that counts the 12,288 system clocks. The input reference for short code I and Q PN generators **202, 206** and long code User PN generator **204** of FIG. 10 in searcher front end **174** is taken from system time count **342**. (Long code User PN generator **204** output is also based on a longer system wide reference which does not repeat for approximately 50 days. The longer system wide reference is not controlled by the searching process and acts as a preset value. The continuing operation based on the preset value is controlled by system time count **342**.) The addresses for PN sequence buffer **176** and even and odd Walsh chip buffers **168** and **170** are taken from system time count **342**. System time count **342** is latched by latch **328** at the beginning of each time slice. The output of latch **328** is selected via address MUX's **330, 332,** and **334** which provide the write addresses corresponding to the current time slice when these buffers are written at some latter time within the time slice.

Offset accumulator **310** keeps track of the offset of the rake element currently being processed. The starting offset as stored in search offset buffer **308** is loaded into offset accumulator **310** at the beginning of each search window. Offset accumulator **310** is decremented with each rake element. At the end of each search rake that is to be repeated for further accumulations, the number of rake elements per search rake as stored in rake width buffer **312** is added back to the offset accumulator to reference it back to the first offset in the search rake. In this way, the searching process again sweeps across the same search rake for another Walsh symbol accumulation. If the searching process has swept across the current search rake on its final Walsh symbol accumulation then offset accumulator **310** is decremented by one by selection of the "-1" input of repeat rake MUX **304** which produces the offset of the first rake element in the next search rake.

The output of offset accumulator **310** always represents the offset of the current rake element being processed and thus is used to control data input to despreader **178**. The output of offset accumulator **310** is added by adders **336** and **338** to the intra-time slice timing output of system time count **342** to

generate the address sequence within a time slice corresponding to a rake element. The output of adders **336** and **338** is selected via address MUX's **330** and **332** to provide antenna sample buffer **172** read addresses.

The output of offset accumulator **310** is also compared by comparator **326** with the output of system time count **342** to form the offset Walsh symbol strobe which indicates that antenna sample buffer **172** has sufficient valid data for the searching process to begin.

Search rake count **320** keeps track of the number of rake elements remaining to be processed in the current search rake. Search rake count **320** is loaded with the width of the search window as stored in search width buffer **314** at the beginning of a search window. Search rake count **320** is incremented after the processing of the final Walsh symbol accumulation of each search rake is complete. When it reaches its terminal count, all offsets in the search window have been processed. To provide a indication that the end of the current search window is imminent, the output of search rake count **320** is summed by summer **324** with the output of rake width buffer **312**. The end of the search window indication marks the time at which antenna sample buffer **172** may begin to be filled with data samples from an alternative antenna in preparation for the next search window without disrupting the contents needed for the current search window.

When channel element microprocessor **136** specifies a search window, it can specify that the search window be performed for a plurality of antennas. In such a case, the identical search window parameters are repeated using samples from a series of antennas. Such a group of search windows is called an antenna search set. If an antenna search set is specified by channel element microprocessor **136**, the antenna set is programmed by the value stored in antenna select buffer **348**. After the completion of an antenna search set, channel element microprocessor **136** is alerted.

Rake element count **318** contains the number of rake elements left to process in the current search rake. Rake element count **318** is incremented once for each rake element processed and is loaded with the output of rake width buffer **312** when the searching process is in the idle state or upon completion of a search rake.

Walsh symbol accumulation count **322** counts the number of Walsh symbols left to accumulate for the current search rake. The counter is loaded with the number of Walsh symbols to accumulate as stored in Walsh symbol accumulation buffer **316** when the searching process is in the idle state or after completing a search rake sweep on the final Walsh symbol accumulation.

Otherwise the counter is incremented with the completion of each search rake.

Input valid count **302** is loaded whenever the input antenna or decimator alignment changes. It is loaded with the minimum number of samples the searching process needs to process a search rake based on the output of rake width buffer **312** (i.e. one Walsh symbol plus one rake width worth of samples). Each time an antenna sample is written to antenna sample buffer **172**, input valid count **302** is incremented. When it reaches its terminal count, it sends an enable signal that allows the searching process to begin. Input valid count **302** also provides the mechanism for holding off search processing when the offsets of successive search windows do not allow continuous processing of data.

The searching process operates in either an idle state, a sync state, or an active state. Searcher sequencing control **350** maintains the current state. Integrated search processor **128** initializes to the idle state when a reset is applied to channel element modem **110**. During the idle state, all counters and accumulators in search control block **166** load their associated search parameters as presented above. Once channel element microprocessor **136** commands the searching process to begin a continuous or a single step search via the control word, integrated search processor **128** moves to the sync state.

In the sync state, the searching process is always waiting for an offset Walsh symbol boundary. If the data in antenna sample buffer **172** isn't valid yet, or if the power control group alignment bit is set and the Walsh symbol is not a power control group boundary, then integrated search processor **128** remains in the sync state until the proper conditions are met on a subsequent offset Walsh symbol boundary. With a properly enabled offset Walsh symbol, the searching process can move to the active state.

Integrated search processor **128** stays in the active state until it has processed a search rake, at which time it normally returns to the sync state. If integrated search processor **128** is in single step mode, it can go from the active state to the idle state after completing the last rake element for the final Walsh symbol accumulation for the last search rake in the search window. Integrated search processor **128** then waits for channel element microprocessor **136** to initiate another search. If instead, integrated search processor **128** is in continuous mode then at this point it loads the new search parameter set and returns to the sync state to await the offset Walsh symbol at the initial offset to be processed in the new search. The active state is the only state in which the antenna data samples are processed. In the idle or sync states the searching process simply keeps track of time with system time

count **342** and continues to write into the PN sequence buffer **176** and antenna sample buffer **172** so that when the searching process does move to the active state these buffers are ready to be used.

FIG. 14 is an exemplary timing view of the first Walsh symbol accumulation of the second search rake in a search window such as search rake **196** shown in FIG. 9. The third Walsh symbol as referenced to the zero offset reference system time clock is shown divided into thirty-two time slices. Searcher state **372** changes from sync to active when the offset Walsh symbol boundary indication corresponding to Walsh symbol 3 indicates that antenna sample buffer **172** is ready with valid samples to process at that offset. During the next available time slice, the first rake element of the search rake is processed. The searching process continues to use each time slice to process a rack element as indicated by an "S" in time slices **374** unless demod front end **122** uses FHT processor engine **120** as indicated by an "D" in time slices **374**. The searching process finishes processing every rake element in the rake and returns to the sync state before the next offset Walsh symbol boundary corresponding to Walsh symbol 4. Also shown is search rake count state **362** being incremented during the active state until it reaches the terminal state, indicating the complete search rake has been processed. Offset count state **364** is shown being incremented between each time slice corresponding to a rake element, so that it may be used to derive the sample buffer offset read address during the time slice. Offset count state **364** is pipelined delayed as offset count for best result register **366**. The offset count **368** is incremented on the final Walsh symbol accumulation **370** pass.

Thus, a single integrated searcher processor configuration, by buffering antenna samples and utilizing a time sliced transform processor, can independently sequence through a search as configured by a search parameter set, analyze the results and present a summary report of the best paths to use for demodulation element reassignment. This reduces the searching related workload of the controlling microprocessor so that a less expensive microprocessor can be used, and also reduces the direct IC costs by allowing a complete channel element modem on a single IC.

The general principles described herein can be used in systems using alternative transmission schemes. The discussion above was based on the reception of a reverse link signal where no pilot signal is available. On the forward link of the preferred embodiment, the base station transmits a pilot signal. The pilot signal is a signal having known data thus the FHT process used to determine which data was transmitted is not necessary. In order to embody the present invention, a integrated search processor for receiving a

signal comprising a pilot signal would not contain the FHT processor or maximum detection function. For example FHT processor engine **120** and max detect **160** blocks of FIG. 5 could be replaced with simple accumulator **125** as shown in FIG. 15. The searching operation when a pilot signal is available is analogous to an acquisition mode access channel search operation as described above.

The searching architecture described above can be used to perform searches in a variety of manners. The most efficient search is a linear search. A linear search is performed by linearly searching potential time offsets in order regardless of the probability that the remote unit is transmitting. When searching for a remote unit signal, the base station must know the expected coverage area range. For example a typical base station covers a range of approximately 50 kilometers implying a round trip delay of 350 microseconds or approximately 430 PN chips in the preferred embodiment. Also, in the multipath environment where signals take nondirect paths, the remote unit signal may be delayed as much as twice the direct path propagation implying that searching must be done over a set of nearly 1000 different PN offsets. Once a remote unit's signal has been detected and is being demodulated, the approximate distance of the remote unit is known and the possible PN offsets which need to be searched to ensure that the majority valid multipath signals are detected are greatly reduced.

Within a given search over a power control group, there are three reasons why a signal may not be detected at a given PN offset. First, no signal may be arriving at the given PN offset. A remote unit may provide several multipath signals but the number of multipath signals created is only a very small portion of all the offsets searched. Thus the majority of searched offsets do not produce energy results that exceed the detection threshold precisely because no remote unit signal is present at that offset.

Secondly, the signal may be arriving at the given PN offset but faded during a large portion of the search integration time. As explained above, the multipath characteristic of a radio channel can result in signal fading. Fading is the result of the phasing characteristics of the multipath channel. A fade occurs when multipath vectors are added destructively, yielding a received signal that is smaller than either individual vector. Thus if a signal which is long term valid happens to be in a deep fade at the time the search is made, no signal is available for detection by the searching process.

Thirdly, the signal would have arrived at the given PN offset but for the fact that the transmitter of the remote unit is gated off during the period of time in question. As explained above, in the preferred embodiment the

remote unit produces a bursty signal. The remote unit comprises a variable rate vocoder which produces variable rate frames of data. The data burst randomizer determines during which time periods the remote unit transmits and during which time periods it does not transmit given the data rate of the signal to be transmitted, a remote unit specific identifying number, and the time of day. When operating at less than full rate, the data burst randomizer within the remote unit pseudorandomly distributes the active time periods within the transmission burst. A corresponding data burst randomizer is also included in the base station such that the base station can recreate the pseudorandom distribution based on the time of day and the remote unit specific identifying number but the rate information is not available during the searching process. As noted above, the eighth rate time periods determine a so called worthy group of time periods. In this way, regardless of the data rate of the transmitted signal, each time period corresponding to the worthy group is sure to correspond to a time when the corresponding remote unit was transmitting a signal. During all other time periods, the remote unit may or may not be transmitting depending on the corresponding encoding rate.

When a linear search is specified, in order to obtain valid power measurements, the searching process confines the search integration time (i.e. the number of Walsh accumulations at a single search offset) to begin and end within a single power control group as explained in greater detail above. A search that integrates only within a single power control group is said to be synchronized with the power control group boundaries. If the searching process at a given offset were accumulated without regard to power control group boundaries and the remote unit were transmitting at less than full rate, valid search results corresponding to a power control group where the remote unit's signal is gated on may be summed with noise accumulated during a subsequent power control group that remote unit's signal is gated off. The summation of the search results corresponding to the power control group where the remote unit's signal is gated off corrupt the otherwise valuable results accumulated during the power control that the remote unit's signal is gated on.

One method of searching would be to search only those power control groups corresponding to worthy groups. Even if such worthy group only searching is performed, the searching process and demodulation element assignment process must still be capable of handling the situation in which the energy accumulated does not exceed the detection threshold but in reality a signal is present at the offset due to the unpredictable fading characteristics

of the channel. Therefore a more efficient scheme is to accumulate energy in all power control groups whether or not they correspond to worthy groups. If energy is detected in a search which does not correspond to a worthy group, an additional valid data point is generated over and above what would be generated based on a worthy group only search.

As noted above, a preamble search and a search performed during traffic channel operation are different. When a remote unit initially attempts to access the system, it sends a beacon signal called a preamble using the Walsh zero symbol. Walsh zero symbol is the Walsh symbol which contains all logical zeros instead of half ones and zeroes as described above. When a preamble search is performed, the searcher looks for any remote unit sending a Walsh zero symbol beacon signal on an access channel. In the preferred embodiment, the transmission of the preamble is always full rate and is never gated off. Therefore during a preamble search there is no need for synchronization with the power control group boundaries.

There are many configurations for spread spectrum multiple access communication systems not specifically described herein but with which the present invention is applicable. For example, other encoding and decoding means could be used instead of the Walsh encoding and FHT decoding. The previous description of the preferred embodiments is provided to enable any person skilled in the art to make or use the present invention. The various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without the use of the inventive faculty. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WE CLAIM:

CLAIMS

1. A method of receiving a signal comprised of a group of spread spectrum call signals sharing a common frequency band wherein each of said spread spectrum call signals comprises a series of bits encoded in groups of a fixed length into a series of symbols wherein a series of said symbols are grouped together in a power control group wherein each symbol in a common power control group is transmitted at a common power level and wherein said power control groups are transmitted in bursts, and isolating one of said call signals from among said group to determine a call signal strength at a path delay time offset from a zero offset reference time, said method comprising the steps of:

storing PN sequence data bits in a PN sequence buffer;

storing a first received set of call signal samples in a sample buffer having a limited size;

despreading a first fixed length set of said call signal samples from said sample buffer corresponding to a first path delay time with a first set of PN sequence data bits from said PN sequence buffer to produce a first despread output;

storing a second received set of call signal samples in said sample buffer; and

despreading a second fixed length set of call signal samples from said sample buffer corresponding to a second path delay time with said first set of PN sequence data bits from said PN sequence buffer to produce a second despread output;

wherein said second fixed length set of call signal samples comprises a large number of the same call signal samples as said first fixed length set of call signal samples and wherein the length of said first and second received set of call signal samples is a fraction the fixed length of said first and second fixed length set of call signal samples;

wherein said steps of storing said first and second fixed length set of call signal samples and said steps of despreading said first and second fixed length set of calls signal samples are performed independent of a probability that said one of said call signals is comprises one of said power control groups.

2. A method of receiving a signal comprised of a group of spread spectrum signals sharing a common frequency band and isolating a first signal from among said group of spread spectrum signals to determine a signal strength at a path delay time offset from a zero offset reference time of said first signal wherein said first signal comprises a series of symbols wherein a series of said symbols are grouped together in a symbol set wherein each symbol in a common symbol set is transmitted at a fixed power level wherein successive symbol sets may be transmitted at a variety of signal levels wherein said variety of signal levels includes a zero level wherein transmission of said first signal is gated off, said method comprising the steps of:

searching a first set of call signal samples corresponding to a first symbol set for said first signal at a first offset to produce a first power estimate thereof;

searching a second set of call signal samples corresponding to said first symbol set for said first signal at said first offset to produce a second power estimate thereof;

summing said first and second power estimates to produce a symbol set power level estimate at said first offset;

searching a third set of call signal samples corresponding to a second symbol set for said first signal at a second offset to produce a third power estimate thereof;

searching a fourth set of call signal samples corresponding to said second symbol set for said first signal at said second offset to produce a fourth power estimate thereof; and

summing said third and fourth power estimates to produce a symbol set power level estimate at said second offset;

wherein said first symbol set and said second symbol set correspond to time contiguous symbol sets and wherein said steps of searching are performed continually regardless of said fixed power level.

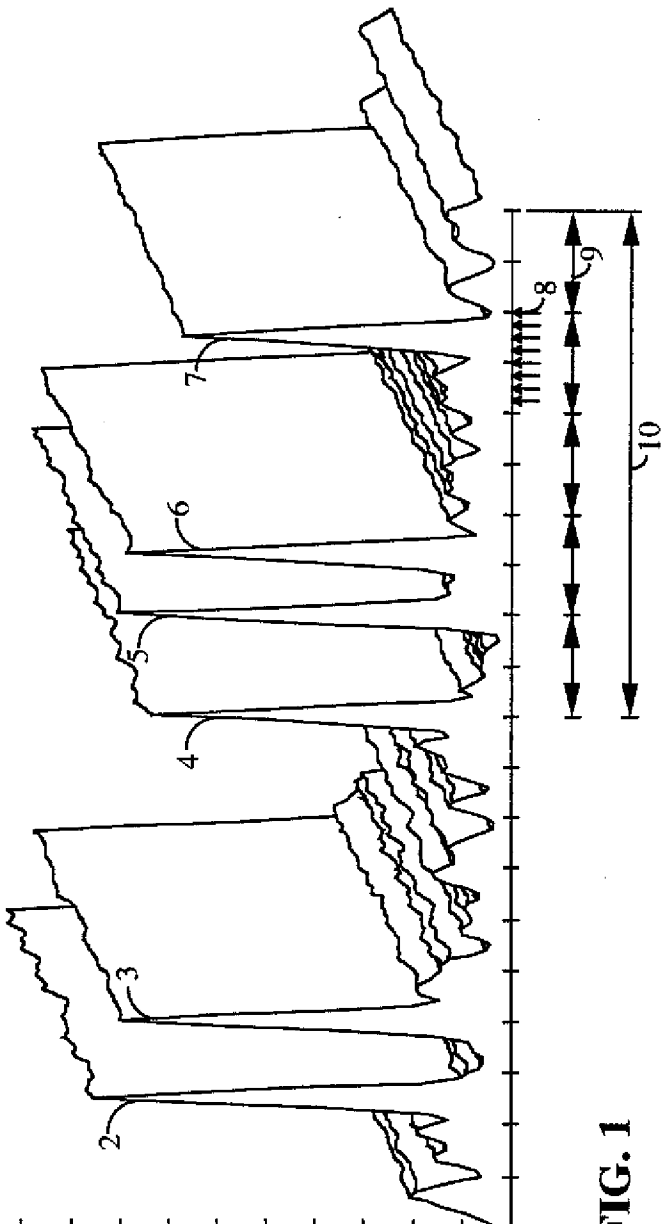
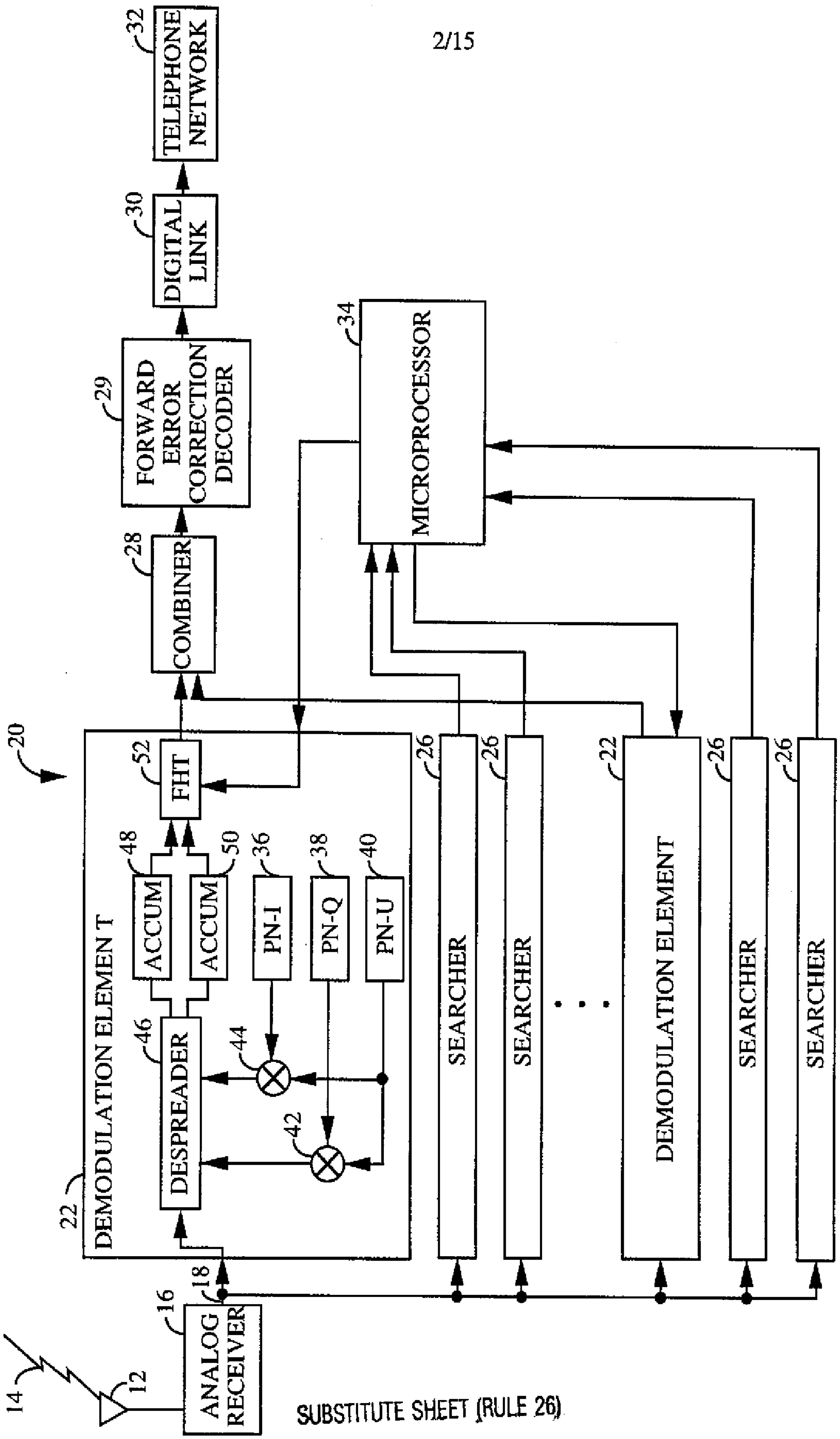


FIG. 1



TELEPHONE NETWORK
32
DIGITAL LINK
30
FORWARD ERROR CORRECTION DECODER
29
COMBINER
28
MICROPROCESSOR
34
20
FHT
52
ACCUM
48
ACCUM
50
PN-I
36
PN-Q
38
PN-U
40
DESPREADER
46
44
42
DEMODULATION ELEMEN T
22
SEARCHER
26
SEARCHER
26
DEMODULATION ELEMENT
22
SEARCHER
26
SEARCHER
26
ANALOG RECEIVER
16
18
14
12

FIG. 2

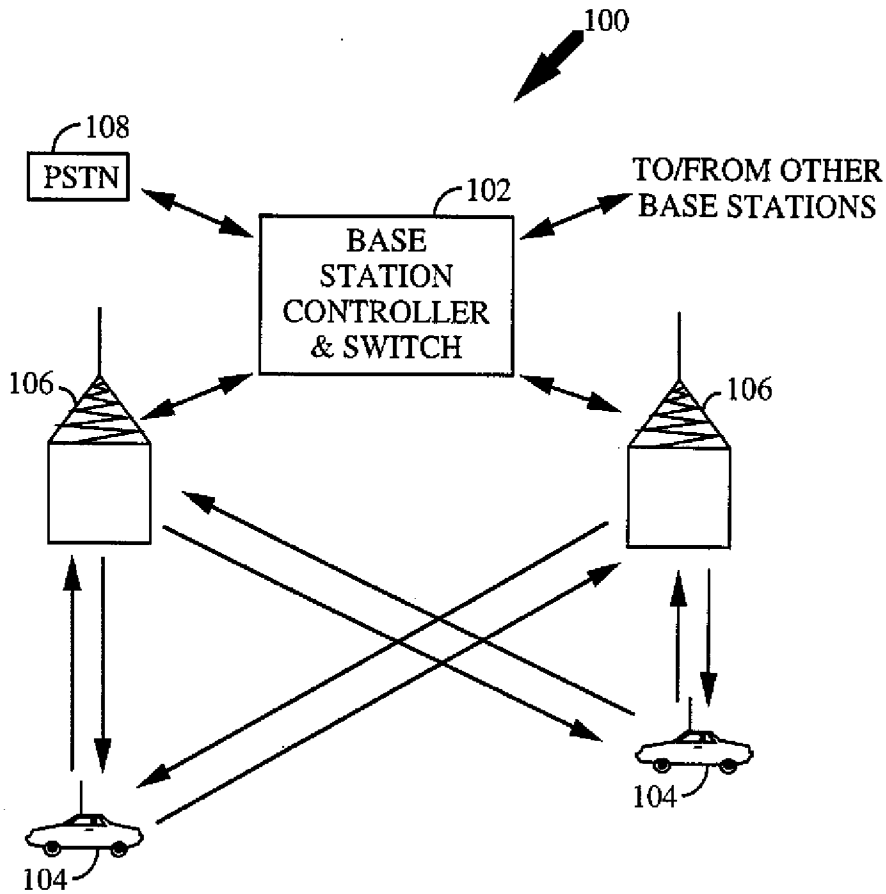

100
108
PSTN
102
BASE STATION CONTROLLER & SWITCH
TO/FROM OTHER BASE STATIONS
106
106
104
104

FIG. 3

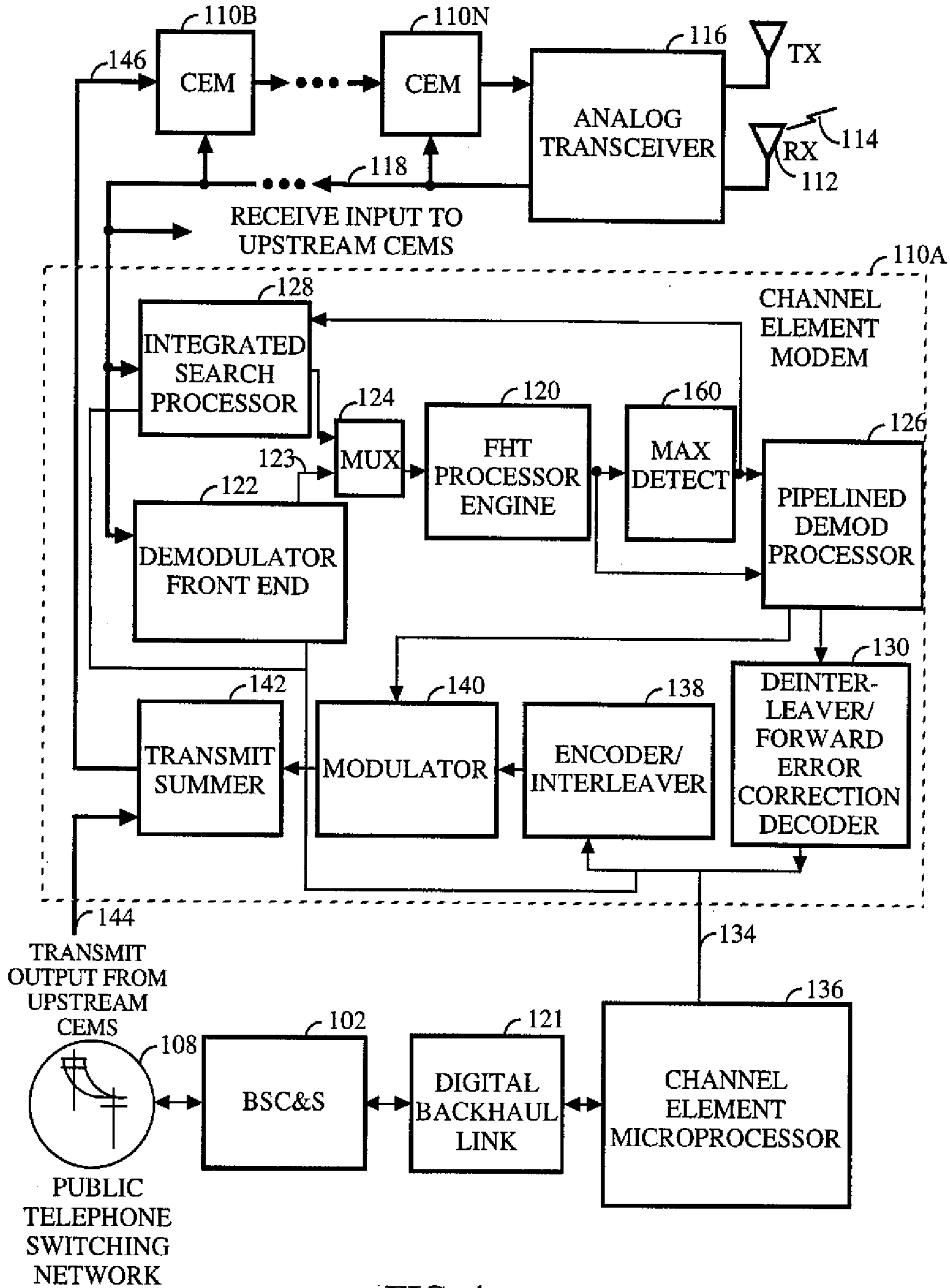

110B
110N
116
TX
146
CEM
CEM
ANALOG
TRANSCEIVER
RX
114
112
118
RECEIVE INPUT TO
UPSTREAM CEMS
110A
CHANNEL
ELEMENT
MODEM
128
INTEGRATED
SEARCH
PROCESSOR
124
MUX
120
FHT
PROCESSOR
ENGINE
160
MAX
DETECT
126
PIPELINED
DEMOD
PROCESSOR
123
122
DEMODULATOR
FRONT END
130
DEINTER-
LEAVER/
FORWARD
ERROR
CORRECTION
DECODER
142
TRANSMIT
SUMMER
140
MODULATOR
138
ENCODER/
INTERLEAVER
144
TRANSMIT
OUTPUT FROM
UPSTREAM
CEMS
134
136
108
102
BSC&S
121
DIGITAL
BACKHAUL
LINK
CHANNEL
ELEMENT
MICROPROCESSOR
PUBLIC
TELEPHONE
SWITCHING
NETWORK

FIG. 4

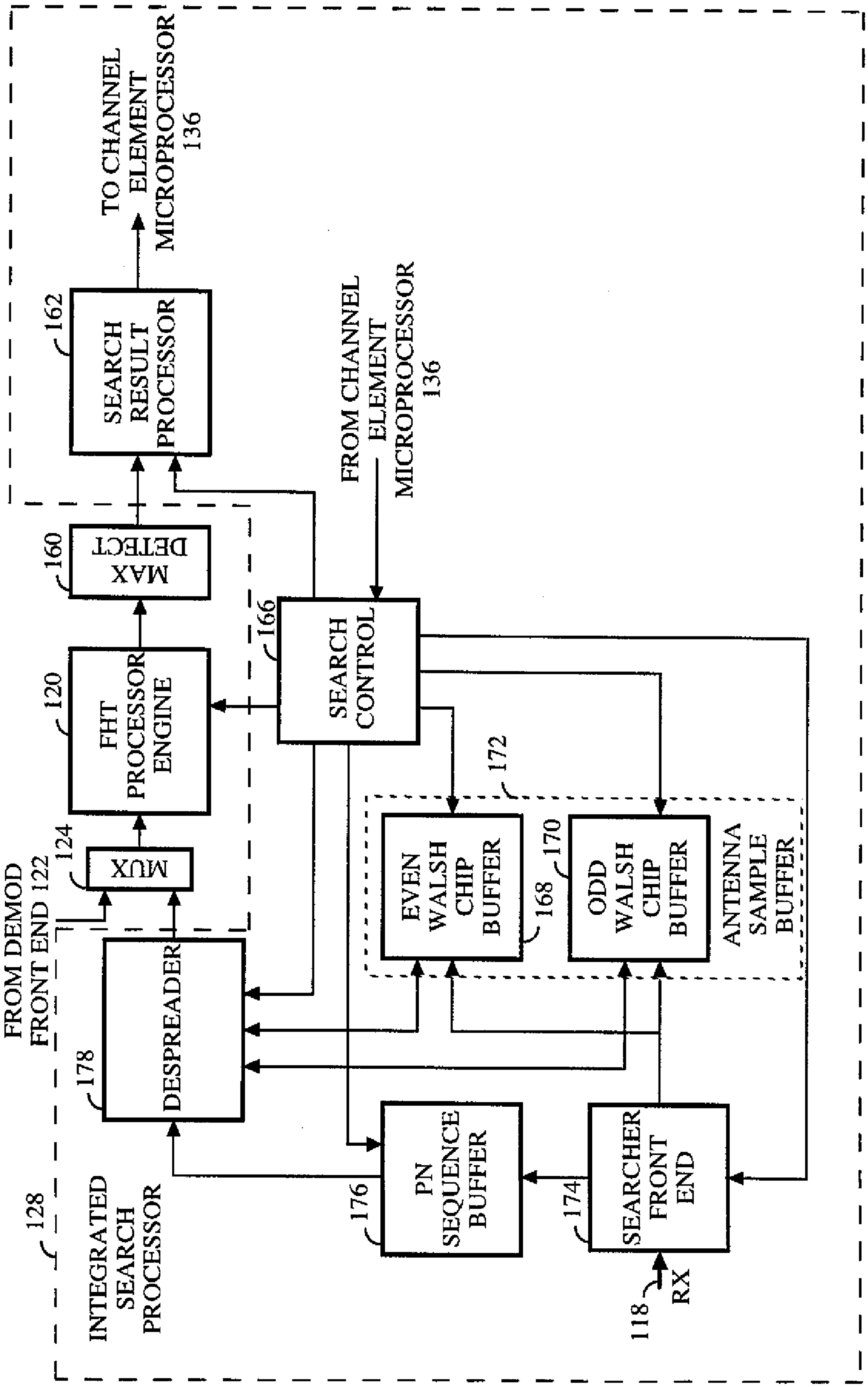

TO CHANNEL ELEMENT MICROPROCESSOR 136
162
SEARCH RESULT PROCESSOR
FROM CHANNEL ELEMENT MICROPROCESSOR 136
160
MAX DETECT
166
SEARCH CONTROL
120
FHT PROCESSOR ENGINE
124
MUX
FROM DEMOD FRONT END 122
172
168
EVEN WALSH CHIP BUFFER
170
ODD WALSH CHIP BUFFER
ANTENNA SAMPLE BUFFER
178
DESPREADER
176
PN SEQUENCE BUFFER
174
SEARCHER FRONT END
118
RX
128
INTEGRATED SEARCH PROCESSOR

FIG. 5

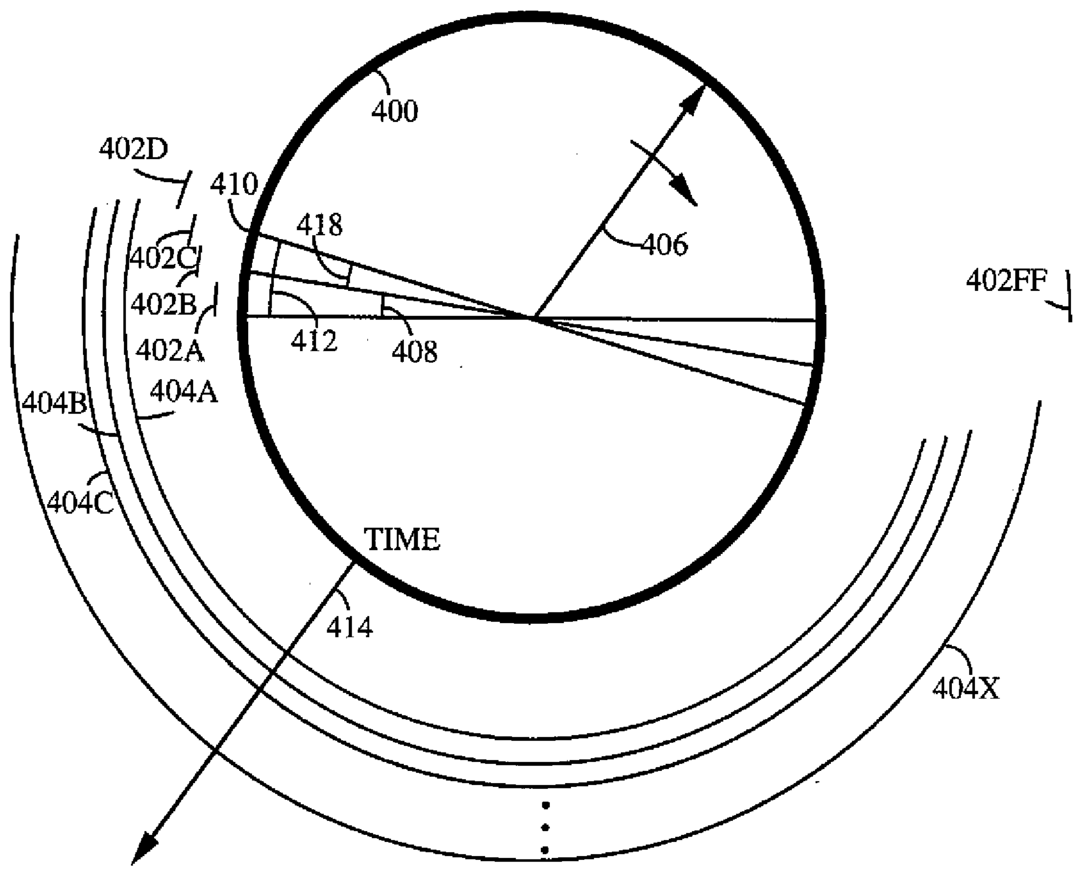

400
402D
410
418
406
402C
402B
402A
412
408
402FF
404B
404A
404C
TIME
414
404X

FIG. 6

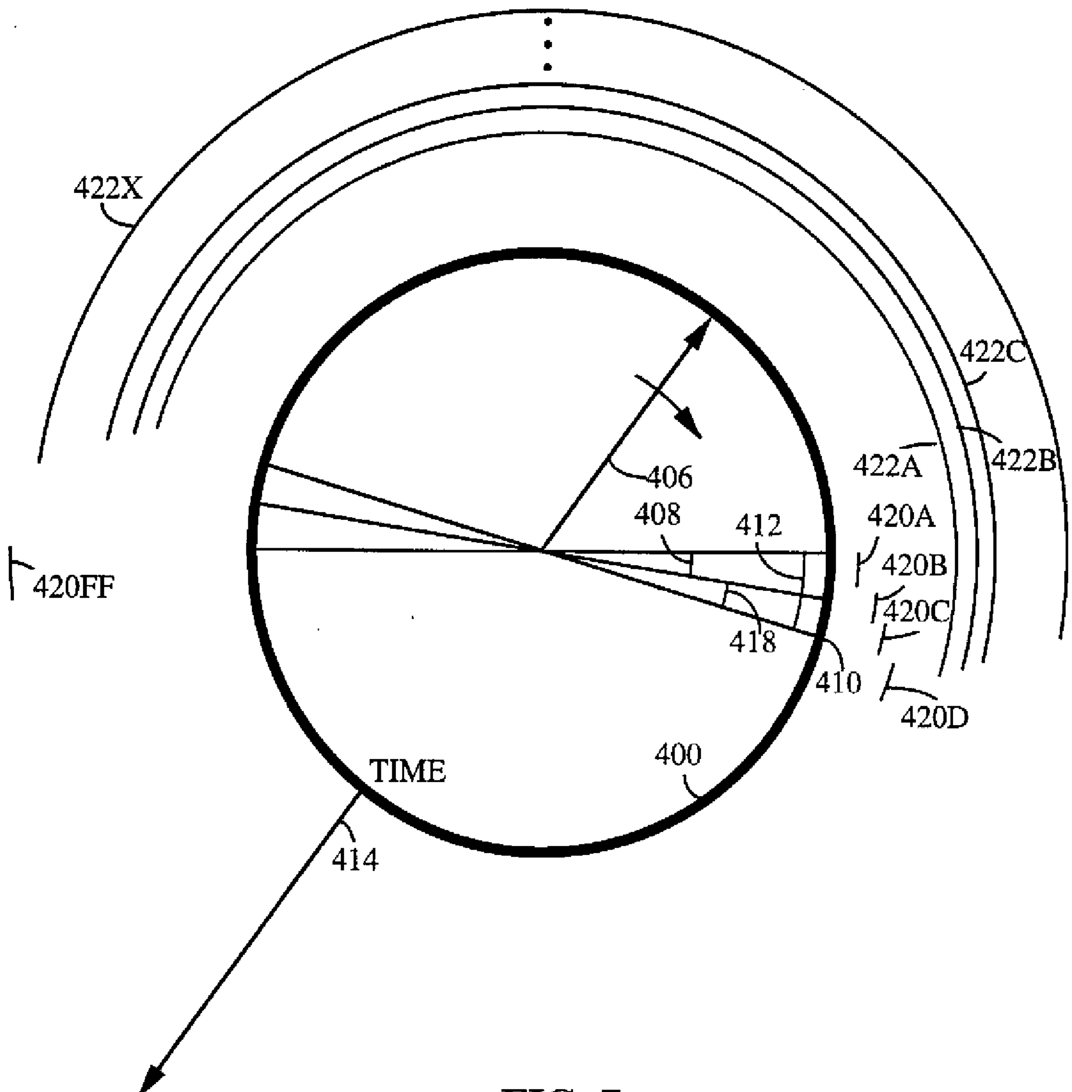

422X
422C
422A
422B
406
408
412
420A
420B
420C
420FF
418
410
420D
TIME
400
414

FIG. 7

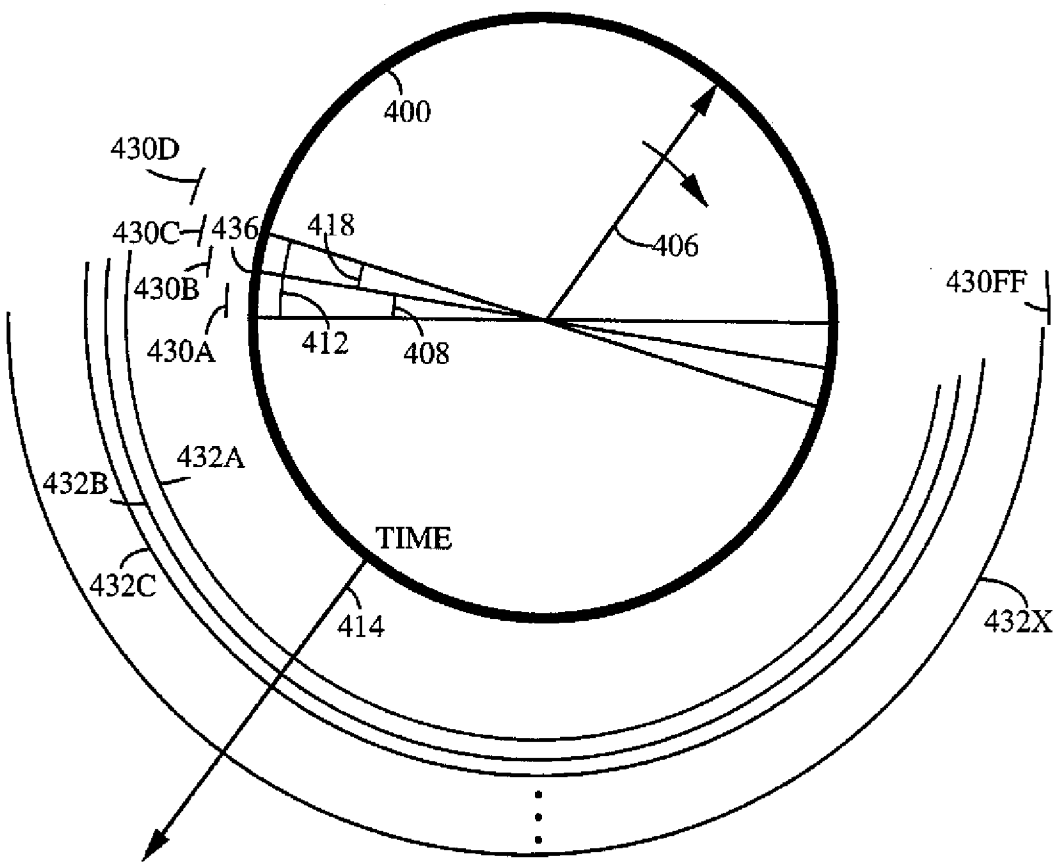

400
430D
430C
436
418
406
430B
430FF
430A
412
408
432A
432B
TIME
432C
414
432X

FIG. 8

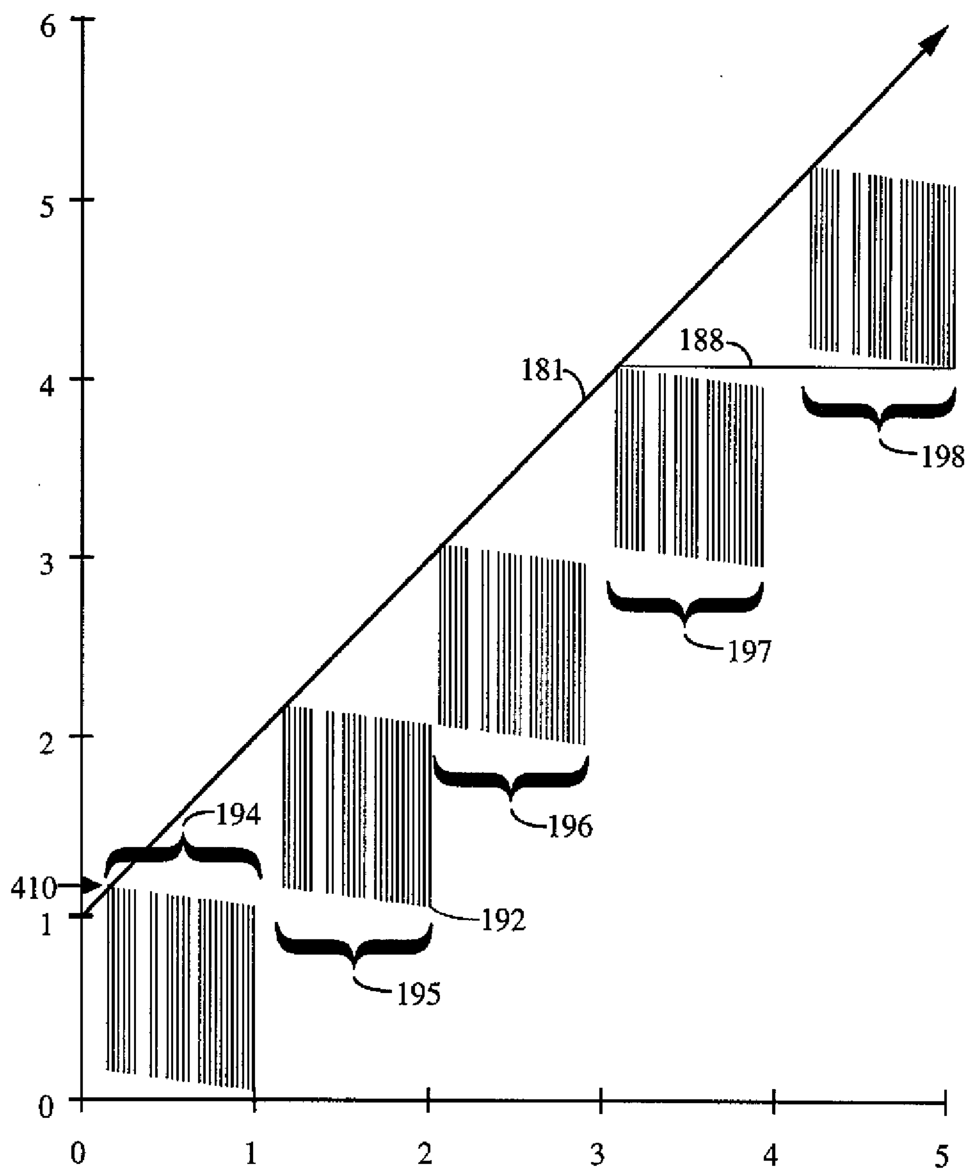


FIG. 9

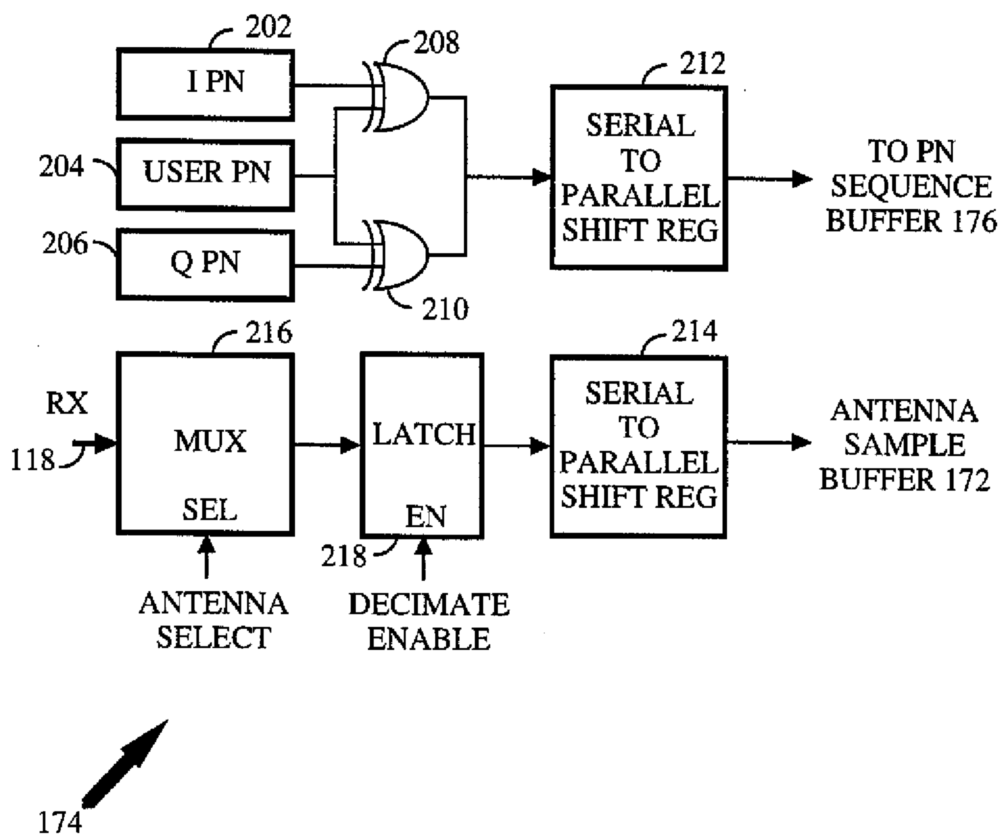
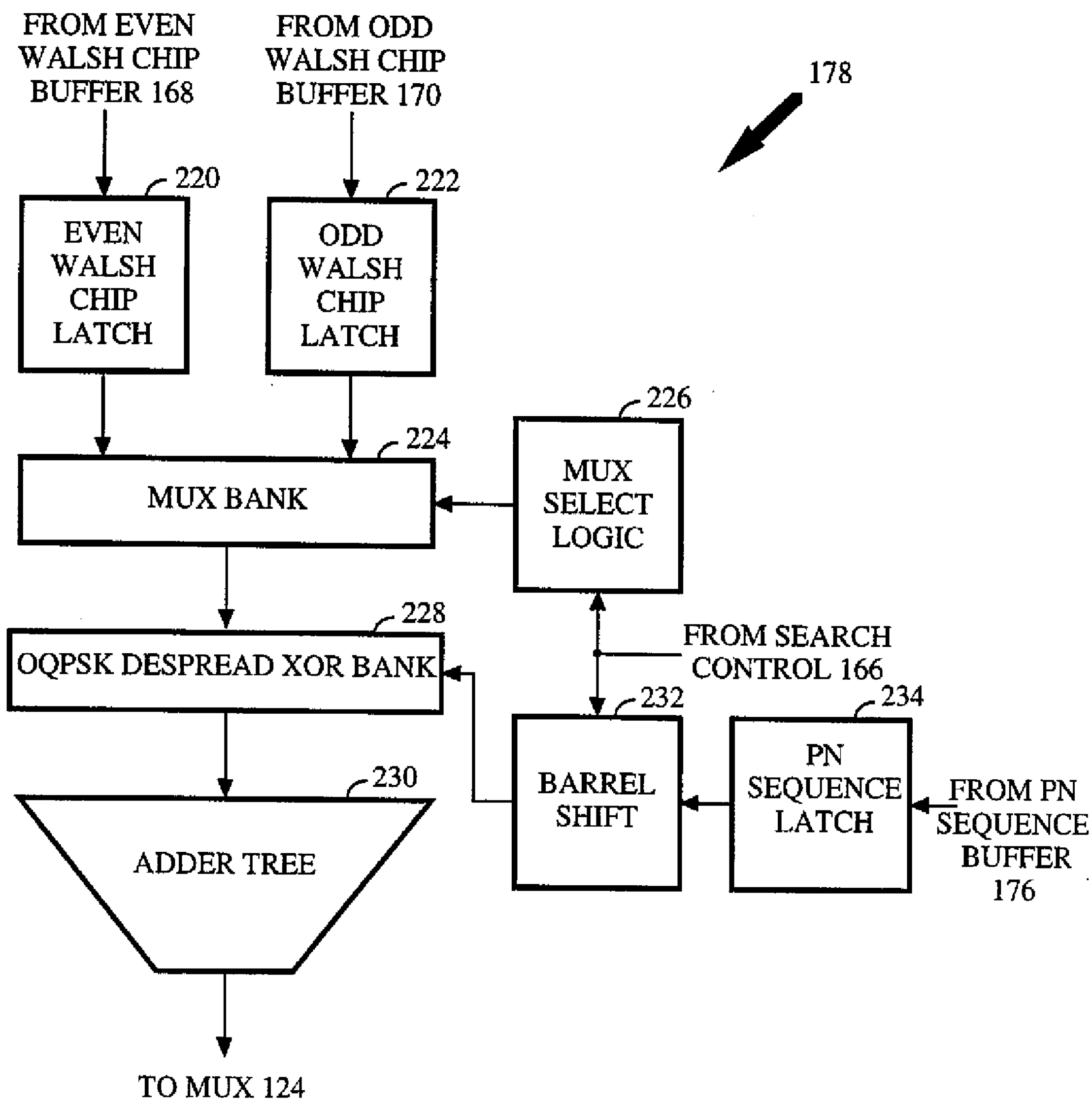

202
I PN
208
212
204
USER PN
SERIAL
TO
PARALLEL
SHIFT REG
TO PN
SEQUENCE
BUFFER 176
206
Q PN
210
216
214
RX
118
MUX
SEL
LATCH
EN
218
SERIAL
TO
PARALLEL
SHIFT REG
ANTENNA
SAMPLE
BUFFER 172
ANTENNA
SELECT
DECIMATE
ENABLE
174

FIG. 10



FROM EVEN WALSH CHIP BUFFER 168
FROM ODD WALSH CHIP BUFFER 170
178
220
222
EVEN WALSH CHIP LATCH
ODD WALSH CHIP LATCH
226
224
MUX BANK
MUX SELECT LOGIC
228
OQPSK DESPREAD XOR BANK
FROM SEARCH CONTROL 166
232
234
230
BARREL SHIFT
PN SEQUENCE LATCH
FROM PN SEQUENCE BUFFER 176
ADDER TREE
TO MUX 124

FIG. 11

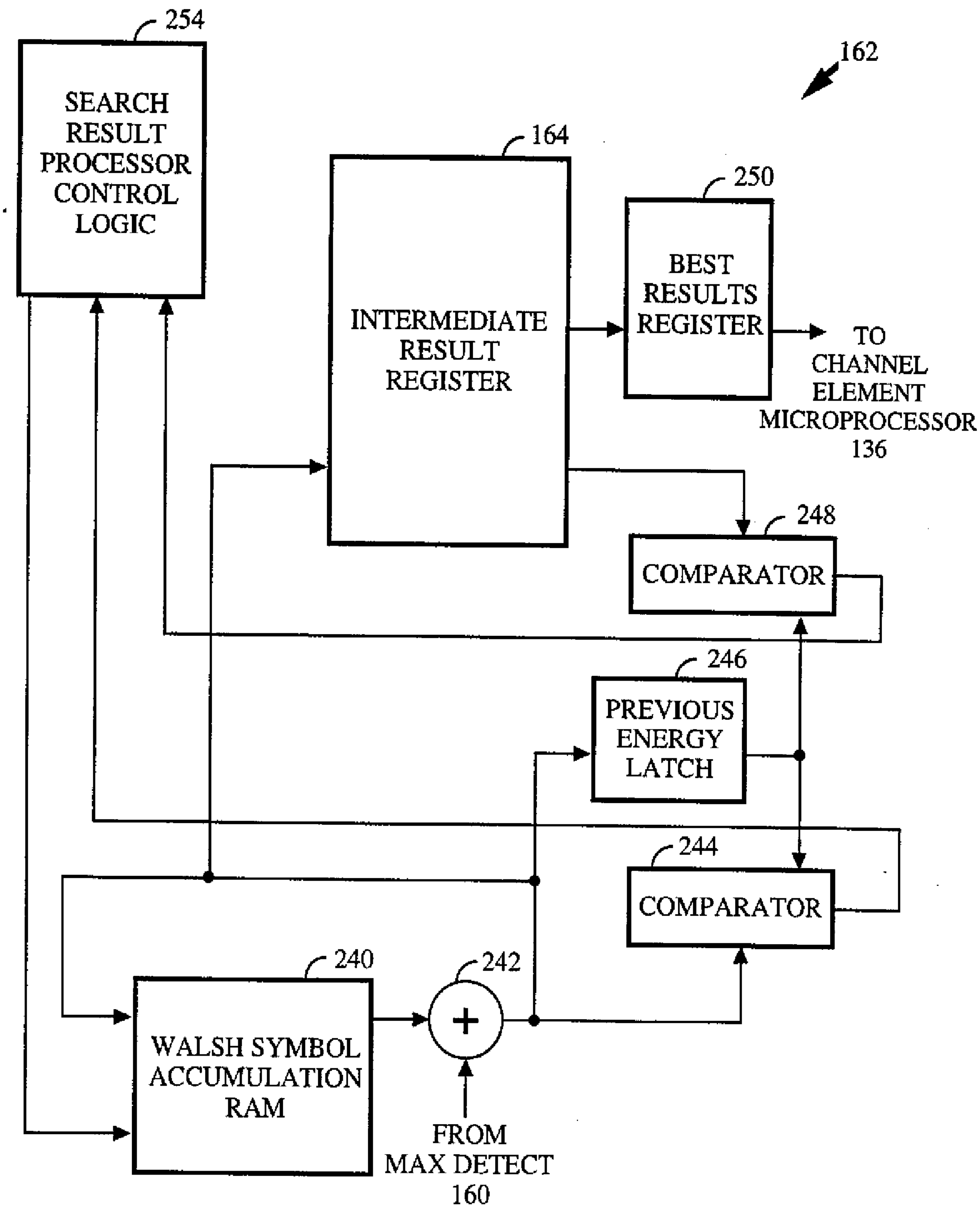

254
SEARCH RESULT PROCESSOR CONTROL LOGIC
162
164
INTERMEDIATE RESULT REGISTER
250
BEST RESULTS REGISTER
TO CHANNEL ELEMENT MICROPROCESSOR 136
248
COMPARATOR
246
PREVIOUS ENERGY LATCH
244
COMPARATOR
240
WALSH SYMBOL ACCUMULATION RAM
242
+
FROM MAX DETECT 160

FIG. 12

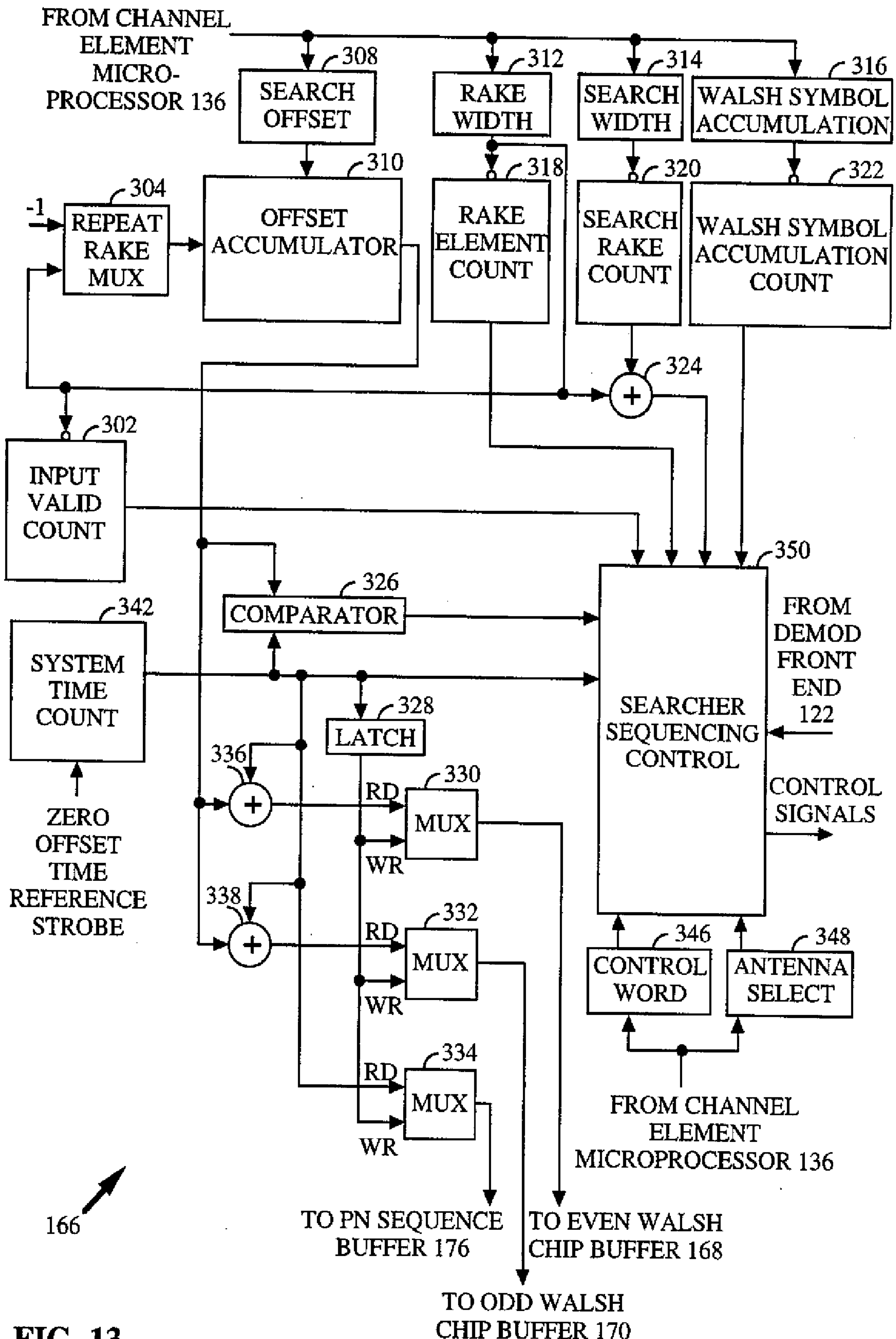
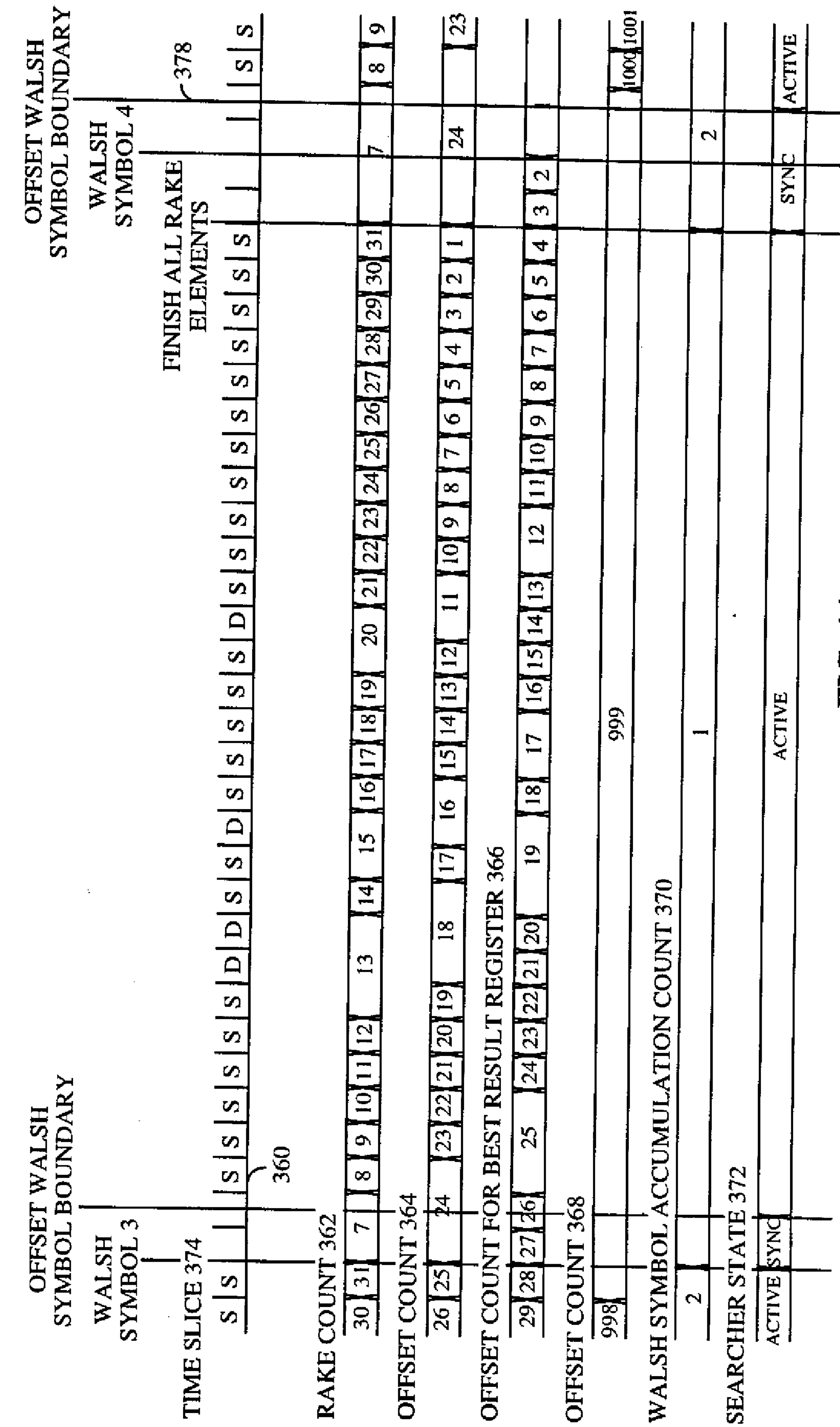

FROM CHANNEL ELEMENT MICRO-PROCESSOR 136
308
SEARCH OFFSET
312
RAKE WIDTH
314
SEARCH WIDTH
316
WALSH SYMBOL ACCUMULATION
310
OFFSET ACCUMULATOR
304
-1
REPEAT RAKE MUX
318
RAKE ELEMENT COUNT
320
SEARCH RAKE COUNT
322
WALSH SYMBOL ACCUMULATION COUNT
324
302
INPUT VALID COUNT
350
326
COMPARATOR
342
SYSTEM TIME COUNT
FROM DEMOD FRONT END 122
328
LATCH
SEARCHER SEQUENCING CONTROL
336
330
RD
MUX
WR
CONTROL SIGNALS
ZERO OFFSET TIME REFERENCE STROBE
338
332
RD
MUX
WR
346
CONTROL WORD
348
ANTENNA SELECT
334
RD
MUX
WR
FROM CHANNEL ELEMENT MICROPROCESSOR 136
166
TO PN SEQUENCE BUFFER 176
TO EVEN WALSH CHIP BUFFER 168
TO ODD WALSH CHIP BUFFER 170

FIG. 13



OFFSET WALSH SYMBOL BOUNDARY
WALSH SYMBOL 3
TIME SLICE 374
360
RAKE COUNT 362
OFFSET COUNT 364
OFFSET COUNT FOR BEST RESULT REGISTER 366
OFFSET COUNT 368
WALSH SYMBOL ACCUMULATION COUNT 370
SEARCHER STATE 372
ACTIVE
SYNC
FINISH ALL RAKE ELEMENTS
WALSH SYMBOL 4
OFFSET WALSH SYMBOL BOUNDARY
378
999
1000
1001

FIG. 14

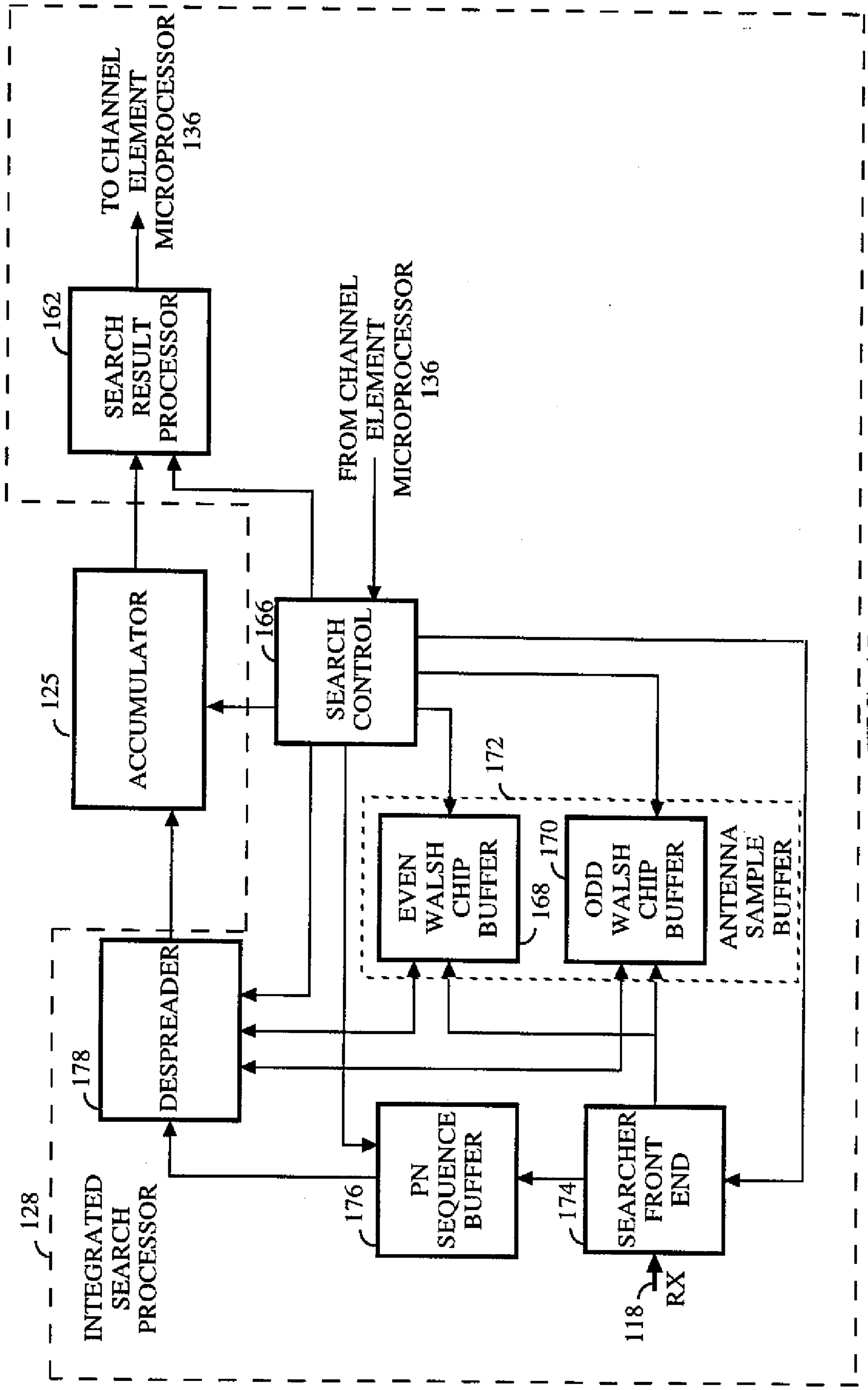

TO CHANNEL ELEMENT MICROPROCESSOR 136
162
SEARCH RESULT PROCESSOR
FROM CHANNEL ELEMENT MICROPROCESSOR 136
125
ACCUMULATOR
166
SEARCH CONTROL
172
EVEN WALSH CHIP BUFFER
168
170
ODD WALSH CHIP BUFFER
ANTENNA SAMPLE BUFFER
178
DESPREADER
128
INTEGRATED SEARCH PROCESSOR
176
PN SEQUENCE BUFFER
174
SEARCHER FRONT END
118
RX

FIG. 15

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 96/07567

A. CLASSIFICATION OF SUBJECT MATTER
IPC 6 H04B7/26 H04B1/707

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 6 H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X,P	WO,A,96 10873 (QUALCOMM INC) 11 April 1996 cited in the application see page 26, line 29 - page 29, line 15; claims 1,16,19-21,24; figures 1,5 ---	1,2
A	WO,A,95 01018 (QUALCOMM INC) 5 January 1995 see abstract; claims 1,2 ---	2
A	US,A,5 109 390 (GILHOUSEN KLEIN S ET AL) 28 April 1992 cited in the application see abstract; figure 3 see column 14, line 47 - column 15, line 40 ---	1,2

-/--

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

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"&" document member of the same patent family

1

Date of the actual completion of the international search	Date of mailing of the international search report
23 September 1996	- 8. 10. 96

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INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 96/07567

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	GB,A,2 278 983 (ROKE MANOR RESEARCH) 14 December 1994 see abstract; claim 1 -----	1,2

1

INTERNATIONAL SEARCH REPORT

Information on patent family members

In tional Application No
PCT/US 96/07567

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
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		CN-A- 1125498	26-06-96
		EP-A- 0705510	10-04-96
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		EP-A- 0653127	17-05-95
		FI-A- 946184	30-12-94
		WO-A- 9428640	08-12-94
		JP-T- 7509831	26-10-95

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60/025,228	29 August 1996 (29.08.96)	US

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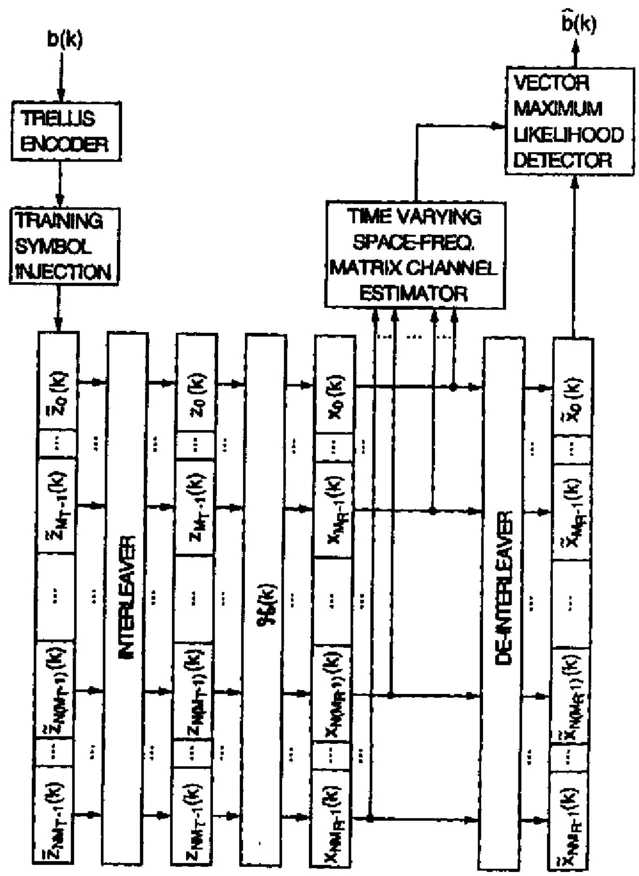
Before the expiration of the time limit for amending the claims and to be republished in the event of the receipt of amendments.

(54) **Title:** HIGH CAPACITY WIRELESS COMMUNICATION USING SPATIO-TEMPORAL CODING

(57) **Abstract**

In a system and method of digital wireless communication between a base station (B) and a subscriber unit (S), a spatial channel characterized by a channel matrix (H) couples an adaptive array of (MT) antenna elements at the base station (B) with an adaptive array of antenna elements (MR) at the subscriber station (S). The method comprises the use of spatio-temporal coding (TRELLIS ENCODER), training symbols (TRAINING SYMBOL INJECTION), and frequency domain deinterleaving (INTERLEAVER). At the receiver, a matched de-interleaver (DE-INTERLEAVER) transforms the space-frequency sequence back into a serial signal stream. A maximum likelihood detector (VECTOR MAXIMUM LIKELIHOOD DETECTOR) generates the recovered information stream.



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High Capacity Wireless Communication
Using Spatio-Temporal Coding

RELATED APPLICATIONS

This application claims priority from U.S. provisional applications 60/025,227 and 60/025,228, both filed 08/29/96. Both applications are hereby incorporated by reference.

FIELD OF THE INVENTION

This invention relates generally to digital wireless communication systems. More particularly, it relates to using antenna arrays by both a base station and a subscriber to significantly increase the capacity of wireless communication systems.

BACKGROUND OF THE INVENTION

Due to the increasing demand for wireless communication, it has become necessary to develop techniques for more efficiently using the allocated frequency bands, i.e. increasing the capacity to communicate information within a limited available bandwidth. This increased capacity can be used to enhance system performance by increasing the number of information channels, by increasing the channel information rates and/or by increasing the channel reliability.

FIG. 1 shows a conventional low capacity wireless communication system. Information is transmitted from a base station B to subscribers $S_1,...,S_9$ by broadcasting omnidirectional signals on one of several predetermined frequency channels. Similarly, the subscribers transmit information back to the base station by broadcasting similar

signals on one of the frequency channels. In this system, multiple users independently access the system through the division of the frequency band into distinct subband frequency channels. This technique is known as frequency division multiple access (FDMA).

A standard technique used by commercial wireless phone systems to increasing capacity is to divide the service region into spatial cells, as shown in FIG. 2. Instead of using just one base station to serve all users in the region, a collection of base stations $B_1, \ldots, B_7$ are used to independently service separate spatial cells. In such a cellular system, multiple users can reuse the same frequency channel without interfering with each other, provided they access the system from different spatial cells. The cellular concept, therefore, is a simple type of spatial division multiple access (SDMA).

In the case of digital communication, additional techniques can be used to increase capacity. A few well known examples are time division multiple access (TDMA) and code division multiple access (CDMA). TDMA allows several users to share a single frequency channel by assigning their data to distinct time slots. CDMA is normally a spread-spectrum technique that does not limit individual signals to narrow frequency channels but spreads them throughout the frequency spectrum of the entire band. Signals sharing the band are distinguished by assigning them different orthogonal digital code sequences. These techniques use digital coding to make more efficient use of the available spectrum.

Wireless systems may also use combinations of the above techniques to increase capacity, e.g. FDMA/CDMA and TDMA/CDMA. Although these and other known techniques increase the capacity of wireless communication systems, there is still a need to further increase system performance. Recently, considerable attention has focused on ways to increasing capacity by further exploiting the spatial domain.

One well-known SDMA technique is to provide the base station with a set of independently controlled directional antennas, thereby dividing the cell into separate sectors, each controlled by a separate antenna. As a result, the frequency reuse in the system can be increased and/or cochannel interference can be reduced. Instead of independently controlled directional antennas, this technique can also be implemented with a coherently controlled antenna array, as shown in FIG. 3. Using a signal processor to control the relative phases of the signals applied to the antenna elements, predetermined beams can be formed in the directions of the separate sectors. Similar signal processing can be used to selectively receive signals only from within the distinct sectors.

In an environment containing a significant number of reflectors (such as buildings), a signal will often follow multiple paths. Because multipath reflections alter the signal directions, the cell space experiences angular mixing and can not be sharply divided into distinct sectors. Multipath can therefore cause cochannel interference between sectors, reducing the benefit of sectoring the cell. In addition, because the separate parts of such a multipath signal can arrive with different phases that destructively interfere, multipath can result in unpredictable signal fading.

In order to avoid the above problems with multipath, more sophisticated SDMA techniques have been proposed. For example, U.S. Pat. No. 5,471,647 and U.S. Pat. No. 5,634,199, both to Gerlach et al., and U.S. Pat. No. 5,592,490 to Barratt et al. disclose wireless communication systems that increase performance by exploiting the spatial domain. In the downlink, the base station determines the spatial channel of each subscriber and uses this channel information to adaptively control its antenna array to form customized beams,

as shown in FIG. 4A. These beams transmit an information signal x over multiple paths so that the signal x arrives to the subscriber with maximum strength. The beams can also be selected to direct nulls to other subscribers so that cochannel interference is reduced. In the uplink, as shown in FIG. 4B, the base station uses the channel information to spatially filter the received signals so that the transmitted signal x' is received with maximum sensitivity and distinguished from the signals transmitted by other subscribers. In this approach the same information signal follows several paths, providing increased spatial redundancy.

In the uplink, there are well known signal processing techniques for estimating the spatial channel from the signals received at the base station antenna array, e.g. by using *a priori* spatial or temporal structures present in the signal, or by blind adaptive estimation. If the uplink and downlink frequencies are the same, then the spatial channel for the downlink is directly related to the spatial channel for the uplink, and the base can use the known uplink channel information to perform transmit beamforming in the downlink. Because the spatial channel is frequency dependent and the uplink and downlink frequencies are often different, the base does not always have sufficient information to derive the downlink spatial channel information. One technique for obtaining downlink channel information is for the subscriber to periodically transmit test signals to the base on the downlink frequency rather than the uplink frequency. Another technique is for the base to transmit test signals and for the subscriber to feedback channel information to the base. If the spatial channel is quickly changing due to the relative movement of the base, the subscriber and/or reflectors in the environment, then the spatial channel must be updated frequently, placing a heavy demand on the system. One method to reduce the required feedback rates is to track only the subspace spanned by the time-averaged channel vector, rather than the instantaneous channel vector. Even with this

reduction, however, the required feedback rates are still a large fraction of the signal information rate.

Although these adaptive beamforming techniques require substantial signal processing and/or large feedback rates to determine the spatial channel in real time, these techniques have the advantage that they can navigate the complex spatial environment and avoid, to some extent, the problems introduced by multipath reflections. As a result, an increase in performance is enjoyed by adaptive antenna array systems, due to their use of the spatial dimension. Note, however, that while the base station antenna array can make efficient use of the spatial dimension by selectively directing the downlink signal to the subscriber S, the uplink signal in these systems is spatially inefficient. Typically, the subscriber is equipped with only a single antenna that radiates signal energy in all directions, potentially causing cochannel interference. These communication systems, therefore, do not make optimal use of the spatial dimension to increase capacity.

OBJECTS AND ADVANTAGES OF THE INVENTION

Accordingly, it is a primary object of the present invention to provide a communication system that significantly increases the capacity and performance of wireless communication systems by taking maximum advantage of the spatial domain. Another object of the invention is to provide computationally efficient coding techniques that make optimal use of the spatial dimensions of the channel. In particular, it is an object of the present invention to provide coding techniques specially adapted for the case of rapidly fading channels where channel state information (CSI) at the transmitter is unknown. These and other objects and advantages will become apparent from the following description and associated drawings.

SUMMARY OF THE INVENTION

These objects and advantages are attained by a method of digital wireless communication that takes maximal advantage of spatial channel dimensions between a base station and a subscriber unit to increase system capacity and performance. Surprisingly, the techniques of the present invention provide an increased information capacity in multipath environments. In contrast, known techniques suffer in the presence of multipath and do not exploit multipath to directly increase system capacity. In brief, the present invention teaches a method of wireless communication using antenna arrays at both the base and subscriber units to transmit distinct information signals over different spatial channels in parallel, thereby multiplying the capacity between the base and the subscriber. In particular, the present invention teaches specific spatio-temporal coding techniques that make optimal use of these additional spatial subchannels in the case of unknown transmitter channel state information.

Generally, the present invention provides a method of digital wireless communication between a base station and a subscriber unit in the case where channel state information is not known by the transmitter. For this purpose a spatio-temporal coding structure that exploits the spatial subchannel capacity is used. In particular, a matrix orthogonal frequency division multiplexing (MOFDM) scheme and a space-frequency trellis coding system is used at the transmitter, and a space-frequency maximum likelihood detector with a channel estimator are used at the receiver. With this relatively simple structure, a MIMO system according to the present invention is able to provide a channel capacity several times greater than can be achieved in a conventional wireless system using OFDM. The inventors also propose an efficient channel estimation algorithm for the time varying MIMO channel.

DESCRIPTION OF THE FIGURES

FIG. 1 shows a low capacity wireless communication system well known in the prior art.

FIG. 2 illustrates a known technique of spatially dividing a service region into cells in order to increase system capacity.

FIG. 3 illustrates the use of beamforming with an antenna array to divide a cell into angular sectors, as is known in the art.

FIGS. 4A and 4B illustrate state-of-the-art techniques using adaptive antenna arrays for downlink and uplink beamforming, respectively.

FIGS. 5A and 5B show the parallel transmission of distinct information signals using spatial subchannels in downlink and uplink, respectively, as taught by the present invention.

FIGS. 6A and 6B are physical and schematic representations, respectively, of a communication channel for a system with multiple transmitting antennas and multiple receiving antennas, according to the present invention.

FIGS. 7A and 7B are block diagrams of the system architecture for communicating information over a multiple-input-multiple-output spatial channel according to the present invention.

DETAILED DESCRIPTION

Although the following detailed description contains many specifics for the purposes of illustration, anyone of ordinary skill in the art will appreciate that many variations and alterations to the following details are within the scope of the invention. Accordingly, the following preferred embodiment of the invention is set forth without any loss of generality to, and without imposing limitations upon, the claimed invention.

As discussed above in relation to FIGS. 4A and 4B, prior art wireless systems employing an adaptive antenna array at the

base station are multiple-input-single-output (MISO) systems, i.e. the channel from the base to the subscriber is characterized by multiple inputs at the transmitting antenna array and a single output at the receiving subscriber antenna. Because these MISO systems can exploit some of the spatial channel, they have an increased capacity as compared to single-input-single-output (SISO) systems that are discussed above in relation to FIGS. 1 and 2. It should be noted that although the MISO systems disclosed in the prior art provide an increase in overall system capacity by spatially isolating separate subscribers from each other, these systems do not provide an increase in the capacity of information transmitted from the base to a single subscriber, or vice versa. As shown in FIGS. 4A and 4B, only one information signal is transmitted between the base and subscriber in both downlink and uplink of a MISO system. Even in the case where the subscriber is provided with an antenna array, the prior art suggests only that this capability would further reduce cochannel interference. Although the overall system capacity could be increased, this would not increase the capacity between the base and a single subscriber.

The present invention, in contrast, is a multiple-input-multiple-output (MIMO) wireless communication system that is distinguished by the fact that it increases the capacity of both uplink and downlink transmissions between a base and a subscriber through a novel use of additional spatial channel dimensions. The present inventors have recognized the possibility of exploiting multiple parallel spatial subchannels between a base station and a subscriber, thereby making use of additional spatial dimensions to increase the capacity of wireless communication. Surprisingly, this technique provides an increased information capacity and performance in multipath environments, a result that is in striking contrast with conventional wisdom.

FIGS. 5A and 5B illustrate a MIMO wireless communication system according to the present invention. As shown in FIG. 5A, a base station B uses adaptive antenna arrays and spatial processing to transmit distinct downlink signals x_1, x_2, x_3 through separate spatial subchannels to a subscriber unit S which uses an adaptive array and spatial processing to receive the separate signals. In a similar manner, the subscriber S uses an adaptive array to transmit distinct uplink signals x'_1, x'_2, x'_3 to the base B over the same spatial subchannels. As the multipath in the environment increases, the channel acquires a richer spatial structure that allows more subchannels to be used for increased capacity.

It is important to note that the simple assignment of the distinct signals to the distinct spatial paths in a one-to-one correspondence, as illustrated above, is only one possible way to exploit the additional capacity provided by the spatial subchannel structure. For example, coding techniques can be used to mix the signal information among the various paths. In addition, the present inventors have developed techniques for coupling these additional spatial dimensions to available temporal and/or frequency dimensions prior to transmission. Although such coupled spatio-temporal coding techniques are more subtle than direct spatial coding alone, they provide better system performance, as will be described in detail below.

It is also important to note that the transmit beamforming at the base requires knowledge of the downlink channel state information. (Similarly, the transmit beamforming at the subscriber requires knowledge of the uplink channel state information. Because the system is symmetric with respect to the base and subscriber, it suffices to discuss one case.) Although downlink channel state information can be fed back to the base from the subscriber, if the channel is rapidly changing, then the demand on the channel capacity to provide real time channel information and the demand on the signal

processing may make it impractical to implement the system under the assumption that transmit channel state information is available. Accordingly, the inventors have developed an MOFDM coding technique to take advantage of the added spatial subchannels even in the case of unknown transmitter channel state information.

In order to facilitate an understanding of the present invention and enable those skilled in the art to practice it, the following description includes a teaching of the general principles of the invention, as well as implementation details. First we develop a compact model for understanding frequency dispersive, spatially selective wireless MIMO channels in the case where the channels are time invariant, and then generalize to the case where the channels vary with time. We then discuss the theoretical information capacity limits of these channels, and propose spatio-temporal coding structures that exploit the spatial subchannel capacity in the case of unknown channel state information. In particular, a matrix orthogonal frequency division multiplexing (MOFDM) scheme is described. In a preferred embodiment a space-frequency trellis coding system is located at the transmitter, and a space-frequency maximum likelihood detector with a channel estimator are located at the receiver. With this relatively simple structure, a MIMO system according to the present invention is able to provide a channel capacity several times greater than can be achieved in a conventional wireless system using OFDM. The inventors also propose an efficient channel estimation algorithm for the time varying MIMO channel.

In its preferred implementations, the present invention makes use of many techniques and devices well known in the art of adaptive antenna arrays systems and associated digital beamforming signal processing. These techniques and devices are described in detail in U.S. Pat. No. 5,471,647 and U.S. Pat. No. 5,634,199, both to Gerlach et al., and U.S. Pat. No.

5,592,490 to Barratt et al., which are all incorporated herein by reference. In addition, a comprehensive treatment of the present state of the art is given by John Livita and Titus Kwok-Yeung Lo in *Digital Beamforming in Wireless Communications* (Artech House Publishers, 1996). Accordingly, the following detailed description focuses upon the specific signal processing techniques which are required to enable those skilled in the art to practice the present invention.

Consider first a time-invariant communication channel for a system with M_T transmitting antennas at a base B and M_R receiving antennas at a subscriber S, as illustrated in FIGS. 6A and 6B. The channel input at a sample time k can be represented by an M_T dimensional column vector

$$\mathbf{z}(k) = [z_1(k),\ldots,z_{M_T}(k)]^T,$$

and the channel output and noise for sample k can be represented, respectively, by M_R dimensional column vectors

$$\mathbf{x}(k) = [x_1(k),\ldots,x_{M_R}(k)]^T,$$

and

$$\mathbf{n}(k) = [n_1(k),\ldots,n_{M_R}(k)]^T.$$

The communication over the channel $\mathbf{H}$ may then be expressed as a vector equation

$$\mathbf{x}(k) = \mathbf{Hz}(k) + \mathbf{n}(k),$$

where the MIMO channel matrix is

$$\mathbf{H} = \begin{pmatrix} h_{1,1} & \ldots & h_{1,M_T} \\ \vdots & & \vdots \\ h_{M_R,1} & \ldots & h_{M_R,M_T} \end{pmatrix}.$$

Each matrix element h_{ij} represents the SISO channel between the i^{th} receiver antenna and the j^{th} transmitter antenna. Due to the multipath structure of the spatial channel, orthogonal

spatial subchannels can be determined by calculating the independent modes (e.g. eigenvectors) of the channel matrix **H**. These spatial subchannels can then be used to transmit independent signals and increase the capacity of the communication link between the base B and the subscriber S.

In the case where the channel matrix **H** is not fixed in time, but changes, it should be represented as a time-dependent matrix, $\mathbf{H}(k)$. Moreover, because the multipath introduces time delays into the various propagation paths, a spatial decomposition of **H** independent of time will result in temporal mixing of the signals. It is more appropriate, therefore, to perform a more general spatio-temporal analysis of the channel.

Let $\{z_j(n)\}$ be a digital symbol sequence to be transmitted from the j^{th} antenna element, $g(t)$ a pulse shaping function impulse response, and T the symbol period. Then the signal applied to the j^{th} antenna element at time t is given by

$$s_j(t) = \sum_n z_j(n) g(t-nT)$$

The pulse shaping function is typically the convolution of two separate filters, one at the transmitter and one at the receiver. The optimum receiver filter is a matched filter. In practice, the pulse shape is windowed resulting in a finite duration impulse response. We assume synchronous complex baseband sampling with symbol period T. We define n_0 and $(\nu+1)$ to be the maximum lag and length over all paths l for the windowed pulse function sequences $\{g(nT - \tau_l)\}$. To simplify notation, it is assumed that $n_0 = 0$, and the discrete-time notation $g(nT - \tau_l) = g_l(n)$ is adopted.

When a block of N data symbols are transmitted, $N+\nu$ non-zero output samples result. Denoting k as the block index for the k^{th} channel usage, $k(N+\nu)$ is the discrete time index for the

first received sample, and $(k+1)(N+\nu)-1$ is the time index for the last received sample. The composite channel output can now be written as an $M_R\cdot(N+\nu)$ dimensional column vector with all time samples for a given receive antenna appearing in order so that

$$\mathbf{x}(k) = [x_1(k(N+\nu)),\ldots,x_1((k+1)(N+\nu)-1),\ldots,\\ x_{M_R}(k(N+\nu)),\ldots,x_{M_R}((k+1)(N+\nu)-1)]^T,$$

with an identical stacking for the output noise samples $\mathbf{n}(k)$. Similarly, the channel input is an $M_T\cdot N$ dimensional column vector written as

$$\mathbf{z}(k) = [z_1(k(N+\nu)),\ldots,z_1(k(N+\nu)+N-1),\ldots,\\ z_{M_T}(k(N+\nu)),\ldots,z_{M_T}(k(N+\nu)+N-1)]^T,$$

The spatio-temporal communication over the channel $\mathbf{H}(k)$ may then be expressed as a vector equation

$$\mathbf{x}(k) = \mathbf{H}(k)\mathbf{z}(k) + \mathbf{n}(k),$$

where the MIMO time-dependent channel matrix

$$\mathbf{H}(k) = \begin{pmatrix} \mathbf{H}_{1,1}(k) & \ldots & \mathbf{H}_{1,M_T}(k) \\ \vdots & & \vdots \\ \mathbf{H}_{M_R,1}(k) & \ldots & \mathbf{H}_{M_R,M_T}(k) \end{pmatrix}$$

is composed of SISO sub-blocks $\mathbf{H}_{ij}(k)$.

To clearly illustrate the effect of multipath, the channel can be written as the sum over multipath components

$$H(k) = \sum_{l=1}^{L} \begin{bmatrix} a_{R,1}(\theta_{R,l})\mathbf{I} \\ \vdots \\ a_{R,M_R}(\theta_{R,l})\mathbf{I} \end{bmatrix} \beta_l(k)\mathbf{G}_l \left[a_{T,1}(\theta_{T,l})\mathbf{I} \ \ldots \ a_{T,M_T}(\theta_{T,l})\mathbf{I} \right].$$

In this equation, $a_{R,j}(\theta_{R,l})$ is the gain response of the j^{th} receiver array element due to angle of arrival $\theta_{R,l}$ of the l^{th} multipath signal, $a_{T,i}(\theta_{T,l})$ is the gain response of the i^{th} transmitter array element due to angle of departure $\theta_{T,l}$ of the l^{th} multipath signal. $\mathbf{B}_l(k)$ is the diagonal time varying channel fading parameter matrix given by

$$\mathbf{B}_l(k) = \mathrm{diag}\ [\ \beta_l(k(N+v)), \ldots, \beta_l((k+1)(N+v)-1)\ ],$$

and the Toeplitz pulse shaping matrix $\mathbf{G}_l$ is given by

$$\mathbf{G}_l = \begin{bmatrix} g_l(0) & 0 & 0 & 0 & \cdots & 0 \\ \vdots & \ddots & & \vdots & & \vdots \\ g_l(v) & \cdots & g_l(0) & 0 & & 0 \\ 0 & g_l(v) & \cdots & g_l(0) & 0 & 0 \\ \vdots & & \ddots & & \ddots & \\ 0 & & 0 & g_l(v) & \cdots & g_l(0) \\ \vdots & & \vdots & & \ddots & \vdots \\ 0 & \cdots & 0 & 0 & 0 & g_l(v) \end{bmatrix}.$$

We will now discuss the information capacity for the spatio-temporal channel developed above. The following analysis assumes that the noise $\mathbf{n}(k)$ is additive white Gaussian noise (AWGN) with covariance $\sigma^2\mathbf{I}_{v+1}$. Each channel use consists of an N symbol burst transmission and the total average power radiated from all antennas and all time samples is constrained to less than a constant.

Write the singular value decomposition (SVD) of the channel matrix as $\mathbf{H}(k)=\mathbf{V}_H(k)\Lambda_H(k)\mathbf{U}_H^*(k)$, with the n^{th} singular value denoted $\lambda_{H,n}(k)$. Write the spatio-temporal covariance matrix for $\mathbf{z}(k)$ for block index k as $\mathbf{R}_Z(k)$ with eigenvalue decomposition $\mathbf{R}_Z(k)=\mathbf{V}_Z(k)\Lambda_Z(k)\mathbf{U}_Z^*(k)$, and eigenvalues $\lambda_{Z,n}(k)$.

It can be demonstrated that, if the case where the instantaneous channel state information is known at both the transmitter and receiver, then the information capacity for

the time-varying discrete-time spatio-temporal communication channel defined above is given by

$$C = E\left(\sum_{n=1}^{K} \log\left(1 + \frac{\lambda_{Z,n}(k)\,|\lambda_{H,n}(k)|^2}{\sigma^2}\right)\right),$$

where $\lambda_{Z,n}(k)$ is given by the spatio-temporal water-filling solution, $E(\cdot)$ is the expectation operator, and K is the number of finite amplitude singular values in $\mathbf{H}(k)$.

For the case where only the receiver has instantaneous channel state information, it is not possible to adapt the transmitter for each block. Nevertheless, it is possible to find the time invariant transmitter covariance which maximizes the capacity for the worst case channel possibilities. For any given transmitted signal covariance matrix $\mathbf{R}_Z$, the worst case channel would place all of the time average energy in the rank 1 subspace defined by the smallest eigendirection in $\mathbf{R}_Z$. This game theoretic problem leads to a spatially uncorrelated transmitter covariance solution $\mathbf{R}_Z = \frac{P_T}{M_T}\mathbf{I}_{M_T}$, where P_T is the maximum average block transmission power. This transmitter covariance is used for completely unknown point to point channels and broadcast channels. For this case of unknown CSI at the transmitter, it can be shown that a white space-time transmission distribution gives a channel capacity

$$C = E\left(\sum_{n=1}^{K} \log\left(1 + \frac{P_T\,|\lambda_{H,n}(k)|^2}{M_T\sigma^2}\right)\right).$$

By analyzing the ranks of the matrices in the path decomposition of the time varying channel $\mathbf{H}(t)$, it can be demonstrated that the maximum number of finite amplitude parallel spatio-temporal channel dimensions, K, that can be created to communicate over the far field time-varying channel

defined above is equal to min{ $N \cdot L$, $(N+\nu) \cdot M_R$, $N \cdot M_T$ }, where L is the number of multipath components. Thus, multipath is an advantage in far-field MIMO channels. If the multipath is large ($L \gg 1$), the capacity can be multiplied by adding antennas to both sides of the radio link. This capacity improvement occurs with no penalty in average radiated power or frequency bandwidth because the number of parallel channel dimensions is increased. In practice, an adaptive antenna array base station, such as that described by Barratt et al., is modified to implement a coding scheme, as described below, which exploits these additional dimensions. In particular, a signal processor is designed to perform a spatio-temporal transform of information signals in accordance with the above equations so that they may be transmitted through the independent parallel subchannels and decoded by the subscriber.

In constant or slowly varying channels, it is often possible to send training sequences to the receiver and communicate channel state information (CSI) back to the transmitter in a manner that accurately tracks time variations. In such cases, the transmitter can implement a coding solution which approaches the theoretical capacity limits. The MIMO communication problem becomes more difficult when the channel fades rapidly in time as is the case with portable wireless communication in the microwave frequency bands. It then becomes impractical to feed back CSI from the receiver to the transmitter due to the information bandwidth required to update the channel state in real time. It is highly desirable in such cases to have a channel coding technique that exploits the spatial dimension of the MIMO problem without requiring any CSI at the transmitter. Such a coding technique has been devised by the present inventors and is described in detail below.

Given the time-varying channel defined by $\mathbf{H}(t)$, it is theoretically possible to create a coding system consisting of

a spatio-temporal encoder, and a spatio-temporal maximum likelihood decoder. The obvious difficulty with such a system is the complexity of the decoder. The complexity of the spatio-temporal decoder can be greatly reduced, however, by using a matrix orthogonal frequency division multiplexing (MOFDM) structure according to the present invention. The complexity reduction occurs because inter-symbol interference (ISI) is eliminated from each OFDM sub-channel.

The MOFDM channel structure is derived under the assumption that the channel is block time invariant over a block of N+2ν symbol periods. Under this assumption, the channel fading matrix $\mathfrak{B}_l(k)$ can be replaced by the scalar fading variable $\beta_l(k)$. Note that the block time invariant assumption is reasonable provided that the block duration $(N+2\nu)T \ll \Delta_\beta$, where Δ_β is the correlation interval for the channel fading variable. (The correlation interval is defined here as the time period required for the fading parameter time-autocorrelation function to decrese to some fraction of the zero-shift value.)

For MOFDM, N data symbols are transmitted during each channel usage. However, a cyclic prefix is added to the data so that the last ν data symbols form a preample to the N data symbol message block. By discarding the first and last ν data symbols at the receiver and retaining only N time samples at the channel output, the new MIMO channel $\hat{\mathbf{H}}(k)$ has a block cyclic structure:

$$\hat{\mathbf{H}}(k) = \sum_{l=1}^{L} \beta_l(k) \begin{bmatrix} a_{R,1}(\theta_{R,l})\mathbf{I} \\ \vdots \\ a_{R,M_R}(\theta_{R,l})\mathbf{I} \end{bmatrix} \hat{\mathbf{G}}_l \left[a_{T,1}(\theta_{T,l})\mathbf{I} \; \dots \; a_{T,M_T}(\theta_{T,l})\mathbf{I} \right].$$

where the cyclic pulse shaping matrix $\hat{\mathbf{G}}(k)$ is given by

$$\hat{\mathbf{G}}_l = \begin{bmatrix} g_l(0) & 0 & \cdots & 0 & g_l(v) & \cdots & g_l(1) \\ \vdots & \ddots & \ddots & & \ddots & \ddots & \\ g_l(v-1) & \cdots & g_l(0) & 0 & \cdots & 0 & g_l(v) \\ g_l(v) & g_l(v-1) & \cdots & g_l(0) & 0 & \cdots & 0 \\ & & & \ddots & \ddots & & \\ 0 & \cdots & 0 & g_l(v) & g_l(v-1) & \cdots & g_l(0) \end{bmatrix}.$$

The MOFDM channel model can now be derived as follows. First post multiply $\hat{\mathbf{H}}(k)$ with the $N \cdot M_T \times N \cdot M_T$ block diagonal inverse discrete Fourier transform (IDFT) matrix $\mathbf{F}^{*(M_T)}$ where each diagonal block is the unitary $N \times N$ IDFT matrix $\mathbf{F}^*$. The next step is to premultiply by a similar $N \cdot M_R \times N \cdot M_R$ block diagonal DFT matrix $\mathbf{F}^{(M_R)}$ where the diagonal submatrices $\mathbf{F}$ are $N \times N$ DFT matrices. Pre- and post-multiplication by permutation matrices $\mathbf{P}_R$ and $\mathbf{P}_T$ then gives the decomposition of the channel into discrete discrete Fourier transform (DFT) frequency domain sub-channels $\mathcal{H}_n(k)$, as follows:

$$\begin{aligned} \mathcal{H}(k) &= \sqrt{(N)}\,\mathbf{P}_R \mathbf{F}^{(M_R)} \hat{\mathbf{H}}(k) \mathbf{F}^{*(M_T)} \mathbf{P}_T \\ &= \begin{pmatrix} \mathcal{H}_1(k) & & 0 \\ & \ddots & \\ 0 & & \mathcal{H}_N(k) \end{pmatrix} \end{aligned}$$

Each channel $\mathcal{H}_n(k)$ is independent of the other frequency domain sub-channels. Just as in the case of scalar OFDM, the cyclic prefix allows the large time domain channel to be decomposed into many smaller parallel frequency domain channels. The received vector signal $\mathcal{X}_n(k)$ for each frequency domain spatial sub-channel can then be expressed as

$$\mathcal{X}_n(k) = \mathcal{H}_n(k)\mathcal{Z}_n(k) + \mathcal{N}_n(k),$$

where $\mathcal{Z}_n(k)$ is the subchannel transmitted signal and $\mathcal{N}_n(k)$ is the subchannel noise. A system architecture implementing this channel structure is shown in FIG. 7A.

The spatial sub-channels can also be expressed as

$$\mathfrak{H}_n(k) = \sum_{l=1}^{L} \beta_l(k)\ \mathfrak{g}_{l,n}\ \mathbf{a}_{R,l}\ \mathbf{a}_{T,l}^T$$

where $\mathfrak{g}_{l,n}$ is the DFT of the sequence $\{g_l(k)\}$ evaluated at DFT index n. At each frequency index, the DMMT channel is due to a weighted sum over L rank-1 outer products of the frequency-invariant receive and transmit array response vectors. The weighting is determined by the frequency invariant path fading values and the Fourier transform of the delayed pulse shaping function. This reveals a highly structured nature for the time varying space-frequency channel spectrum.

In the case of rapidly fading channels where CSI is not available at the transmitter, the appropriate transmitter distribution is a spatially and temporally white transmitter sequence. Nevertheless, as seen from the above channel decomposition, the use of cyclic signal structures allows the determination of a channel structure that can still be exploited to improve capacity. Therefore, a practical subchannel coding method which approximates a white distribution is desired. Although many variations are possible, the following description is focused on a particularly simple strategy involving a one dimensional trellis coding structure.

In MOFDM, a space-frequency code is transmitted. Given M_T transmitting antenna elements and the MOFDM subchannel decomposition of $\mathfrak{H}(k)$, a codeword sequence $\mathbf{c}^{(j)}$ of constraint length $N_c^{(j)}$ can be viewed as q spatial vector code segments transmitted in each of $q = \frac{|\mathbf{c}^{(j)}|}{M_T}$ frequency bins where $|\mathbf{c}^{(j)}|$ is the length of the code sequence. In this embodiment, an information signal b(k) is converted into a code sequence $\mathbf{c}^{(j)}$ by a one dimensional trellis encoder, as shown in FIG. 8. Code

segments of length M_T form a spatial vector code $\mathbf{c}_n^{(j)}$ for a single MOFDM frequency bin indexed by n. After training symbols are injected, frequency domain interleaving is performed by an interleaver in order to distribute consecutive spatial vector code segments among well separated frequency bins. Interleaving allows the system to exploit the frequency diversity of the channel while the spatial coding is a form of spatial diversity.

Each of the M_T symbols in a given spatial vector code segment for a given frequency bin are transmitted from one of the antennas. At the receiver, a matched frequency de-interleaver transforms the space-frequency sequence back into a serial signal stream. A tilde, ~, above a variable is used to denote the signal sequence before interleaving and after de-interleaving operations. Define

$$\mathbf{c}^{(j)} = [\ c_0^{(j)}, \ldots, c_{qM_T-1}^{(j)}\ ]^T$$

as the trellis encoder symbol sequence codeword of length qM_T indexed by j. Further define

$$\tilde{\mathfrak{X}}^{(q)}(k) = [\ \tilde{\mathfrak{X}}_{\ell M_R}(k),\ \ldots,\ \tilde{\mathfrak{X}}_{\ell M_R+qM_T-1}(k)\ ]^T$$

as the received de-interleaved signal sequence due to the transmitted code $\mathbf{c}^{(j)}$ where ℓM_R is the beginning index for the received sequence of length qM_R spanning q space-frequency subchannels. The output sequence due to codeword $\mathbf{c}^{(j)}$ can now be written as

$$\tilde{\mathfrak{X}}^{(q)}(k) = \sqrt{\frac{P_T}{M_T}}\ \tilde{\mathfrak{H}}^{(q)}(k)\mathbf{c}^{(j)} + \tilde{\mathfrak{N}}^{(q)}(k)$$

where

$$\mathfrak{H}^{(q)}(k) = \begin{pmatrix} \tilde{\mathfrak{H}}_l(k) & & 0 \\ & \ddots & \\ 0 & & \tilde{\mathfrak{H}}_{l+q}(k) \end{pmatrix}$$

and

$$\sqrt{\frac{P_T}{M_T}}\,\mathbf{c}^{(j)} = \begin{pmatrix} \tilde{\mathfrak{Z}}_{lM_T}(k) \\ \vdots \\ \tilde{\mathfrak{Z}}_{l+qM_T-1}(k) \end{pmatrix}.$$

The additive noise term $\tilde{\mathfrak{N}}^{(q)}(k)$ is still white after the MOFDM channel operations.

For a given spatio-temporal symbol code set

$$\mathbf{C} = \{\mathbf{c}^{(1)}, \ldots, \mathbf{c}^{(J)}\}$$

the maximum likelihood detector is given by

$$\hat{\mathbf{c}} = \arg \max_{\mathbf{c}(j)} P(\mathbf{c}^{(j)} | \tilde{\mathfrak{X}}^{(q)}(k)).$$

FIG. 7B shows such a detector which is used to generate the recovered information stream, $\hat{b}(k)$. Given that the receiver noise present in each space-frequency sub-channel is multivariate AWGN, it is known that the equivalent decoder optimization is

$$\hat{\mathbf{c}} = \arg \min_{\mathbf{c}(j)} \left|\left|\sqrt{\frac{P_T}{M_T}}\,\tilde{\mathfrak{H}}^{(q)}(k)\mathbf{c}^{(j)} - \tilde{\mathfrak{X}}^{(q)}(k) = \right|\right|_2^2.$$

This equation can be solved efficiently using a vector Viterbi detector similar to that used in ISI channels. The main difference here is that while there is correlation in the received spatial code segment in each frequency bin, the information across frequency bins is uncorrelated. This allows the metric computation to be pruned back to the number

of states in the trellis encoder at the beginning of each new spatial code segment hypothesis test. It is undesirable for the encoder to posses parallel transitions because this reduces the diversity order of the code to one. Therefore, all of the encoder input bits are fed to the convolutional encoder with rate r and there are there is only one member in each of the cosets.

The inventors have discovered that, given a random Rayleigh channel process with uncorrelated spatial fading and perfect frequency domain interleaving and any code set **C**, the upper bound on the average bit error rate for a $1 \times m$ SIMO channel is larger than the bound for a $M \times mM$ MIMO channel, even though the later transmits data at M times the rate of the former. This remarkable fact reveals some very interesting behavior for the proposed MIMO channel coding structure. Although the data rate for the $M_R = M = M_T$ MIMO channel goes up linearly with M, the probability of error bound is smaller than that for the SISO channel. While the transmitter power for each spatial symbol must be reduced as the number of antennas and spatial sequences are increased, the length of the spatial code segment error vector $\mathbf{e}_n(j1, j2)$ also increases to offset the transmitter power reduction. In addition, while the frequency diversity due to the number of frequency bins spanned by the code error sequence is reduced as M increases, the denominator exponent increases due to spatial diversity. Thus, as M increases, the effects of frequency diversity are replaced by spatial diversity. Furthermore, it is clear that the MIMO system can benefit from additional spatial diversity by setting $M_R > M_T$. The m-order spatial diversity error performance of a $1 \times m$ SIMO channel can be achieved with an $M \times mM$ MIMO channel which will again achieve M times the data rate of the SIMO channel while maintaining lower error probability.

An alternative code design metric for the spatio-temporal coding structure presented in relation to FIGS. 7A and 7B is

suggested by observing that the correct error sequence metric for code design is clearly the product of Euclidean distances for each of the M_T length spatial error vector segments in the error event sequences. This metric is strikingly similar to periodic product distance metrics that are known from other contexts.

An important aspect of the present invention is channel estimation. In fast fading channels, overhead penalties for conventional multi carrier training techniques can be severe. A large number of sub-channels N is desired so that cyclic prefix overhead is minimized. Large N corresponds to long OFDM symbol duration. Long symbol duration, in turn, requires short intervals between training. In conventional channel training procedures, an entire OFDM symbol is dedicated to training, and several data symbols are inserted between training symbols. Thus, a trade-off exists between cyclic prefix and training overhead with conventional channel estimation techniques.

Furthermore, in burst-mode transmission applications such as wireless ATM, if the average data rate for a virtual circuit is low, then the time between ATM packets can be large. In such cases, it is not feasible to use an entire DMT symbol for training since the channel can change substantially between training symbols. What is needed is a training strategy that allows "instantaneous updates" for the channel estimation algorithm. The present inventors have developed a training approach and channel estimation algorithm which injects training information along with data into each OFDM symbol. The channel estimation algorithm exploits the correlation properties of the time varying wireless channel to estimate the spatial channel for each MOFDM frequency domain sub-channel.

Imperfect channel knowledge can have an impact on error probability, and channel estimation noise in the receiver will

limit the performance of a spatio-temporal coding system. The inventors have discovered that the effect of channel estimation errors can be modeled as an increase in the effective noise variance. This noise variance increase is an interesting function of the time varying channel correlation function, the portable velocity, the average channel SNR and the design of the channel estimation algorithm. Thus, proper design of the channel estimator is critical for low error probability communication.

In all that follows, we again invoke the spatially uncorrelated Rayleigh fading condition. Although the spatial fading is uncorrelated, there is correlation in the OFDM frequency and time domains which we wish to exploit. The correlation in the frequency domain arises from the delay limited nature of the channel impulse response. The correlation in the time domain fading arises from the band limited Doppler shifts experienced by physical objects which move in the vicinity of the portable. We desire a channel estimation algorithm that exploits these correlation properties in an optimal manner.

To estimate the matrix channel that exists at a given OFDM frequency index, note that we can simply estimate the M_T column vectors of dimension $1 \times M_R$. Given that the column vectors are assumed to fade independently, an optimal training strategy is to transmit M_T different training sequences from each transmitter antenna and estimate the resulting column vectors without considering the information received during training from the other transmitter antennas. In addition, the uncorrelated spatial fading assumption allows each of the scalar elements in a given channel column vector to be estimated independently. Thus, with M_T training sequences transmitted independently from each antenna, we can estimate M_R independent frequency domain scalar channel entries. Thus, the focus is on a SISO training strategy for the frequency domain sub-channels that exist between one transmitter antenna

and one receiver antenna. The SISO estimation algorithm is directly generalized to the MIMO case by stacking SISO estimates from each receiver antenna into columns and exploiting the cyclic shift properties of the DFT.

Our SISO channel estimation strategy will be to transmit training symbols in several equally spaced OFDM sub-channels with data embedded between training symbols. For a discrete time channel which is delay limited to $\nu+1$ finite impulse response terms, $\nu+1$ OFDM training sub-channels are sufficient to construct an estimate of all N sub-channels.

A SISO OFDM channel estimation algorithm is now described. The channel frequency domain training symbol sequence is defined as

$$\mathcal{Z}_T = \mathrm{diag}\left[\mathcal{Z}_0, \mathcal{Z}_{\frac{N}{\nu+1}}, \ldots, \mathcal{Z}_{\frac{\nu N}{\nu+1}}\right].$$

By construction, $\mathcal{Z}_T \mathcal{Z}_T^{*} = P_T \mathbf{I}_{\nu+1}$.

The channel estimation procedure is as follows.

1. Given n_1 past measurements, the present measurement, and n_2 future measurements of the frequency-domain training sub-channel outputs, form n_1+n_2+1 measurements of the time varying channel impulse response vector $\tilde{\mathbf{h}}^{(\nu+1)}(k)$ by dividing the received known training symbols into the outputs and then performing the IDFT operation, i.e.

$$\tilde{\mathbf{h}}^{(\nu+1)}(k) = \sqrt{\frac{\nu+1}{N}}\, \mathbf{F}_{\nu+1}^{*} \mathcal{Z}_T^{-1} \mathfrak{X}_T^{\nu+1}(k),$$

where $\mathfrak{X}_T^{\nu+1}(k) = \left[\mathfrak{X}_0(k), \mathfrak{X}_{\frac{N}{\nu+1}}(k), \ldots, \mathfrak{X}_{\frac{\nu N}{\nu+1}}(k)\right]^T$ and $\mathbf{F}_{\nu+1}$ is the $\nu+1$ point DFT matrix.

2. Form the channel impulse response estimate $\hat{\mathbf{h}}^{(v+1)}(k)$ by applying an optimal linear MMSE estimation filter independently to each of the impulse response measurements $\tilde{\mathbf{h}}^{(v+1)}(k)$, i.e.

$$\hat{\mathbf{h}}^{v+1}(k) = \left[\mathbf{w}_h^{*} \otimes \mathbf{I}_{v+1}\right]\begin{bmatrix}\tilde{\mathbf{h}}^{v+1}(k-n_1)\\ \vdots \\ \tilde{\mathbf{h}}^{v+1}(k-n_2)\end{bmatrix},$$

where $\otimes$ denotes the Kronecker product and $\mathbf{w}_h$ is the scalar Weiner filter for $\hat{\mathbf{h}}^{(v+1)}(k)$.

3. Form the complete OFDM channel estimate by zero-padding the channel impulse response estimate and performing an N-point FFT, i.e.

$$\mathfrak{H}^{(N)} = \mathbf{F}_N \left[\ \hat{\mathbf{h}}^{(v+1)}(k),\ 0,\ \ldots,\ 0\ \right]^T.$$

Given the iid fading assumption on the channel impulse response terms, the above channel estimation algorithm is optimal within the class of linear MMSE estimators.

To extend the preceding scalar channel analysis methods to estimate the OFDM matrix subchannels, the following procedure is employed. Rather than transmitting v+1 training symbols spaced by $\frac{N}{v+1}$ sub-channels, we transmit v+1 frequency domain sequences, each of length M_T. This training scheme is illustrated in Table 1 where the notation T(n) represents training symbol n and D(k) represents data symbol k.

Table 1

FFT bin	Content
0	T(0)
:	:
M - 1	T(M-1)
M	D(0)
:	:
N/(ν+1) - 1	D(N/(ν+1) - M - 1)
:	:
νN/(ν+1)	T(νM)
:	:
νN/(ν+1) + M - 1	T(M(ν+1) - 1)
νN/(ν+1) + M	D((ν-1)(N/(ν+1)-M))
:	:
N - 1	D(νN/(ν+1)-νM - 1)

In each of the M_T long training sequences, the first symbol is transmitted from the first antenna, the second symbol from the second antenna, and so on. The first column in the frequency domain sub-channel matrix response is then estimated by performing the scalar channel estimation algorithm on each of the M_R antenna outputs associated with the $\nu+1$ subchannels which appear first in the M_T long training sequences, and stacking the scalar estimates into a vector. The other columns of the matrix channel are estimated in a similar manner, with the exception that the final frequency domain estimates $\hat{\mathfrak{H}}(k)$ obtained from the channel estimation algorithm are cyclic-shifted to account for the frequency sub-channel offset for each transmitter antenna training sequence.

Using this technique, for a complex $M_T \times M_R$ DMT channel, only $M_T(\nu+1)$ DMT sub-channels are required for training. The channel overhead loss due to training and the cyclic prefix is

then $\frac{M_T(v+1)+v}{N+v}$. This training technique only requires FFTs and FIR channel estimation filters to implement.

Among the various applications of the present invention, one of particular utility is a wideband wireless ATM local area network for a campus environment. The transmitted digital symbol rate is 10 MHz. The portable terminals are mobile with a maximum velocity of 70 miles per hour. The RF carrier frequency is 5.2 GHz. This application is extremely challenging for conventional equalizer based communication structures due to the large delay spread and extremely high Doppler frequency (+/- 540 Hz). The Doppler shift also makes conventional CDMA approaches difficult due to the required power control loop bandwidth. For these reasons, the application is an ideal candidate for MOFDM. In one embodiment, such a MOFDM system may have 3 transmitter antennas and either 3 or 6 receiver antennas.

Thus, it will be clear to one skilled in the art that the above embodiment may be altered in many ways without departing from the scope of the invention. Accordingly, the scope of the invention should be determined by the following claims and their legal equivalents.

CLAIMS

What is claimed is:

1. A method of digital wireless communication between a base station and a subscriber unit, the method comprising:

space-frequency encoding a plurality of information signals into a sequence of transmitted signal vectors, wherein the transmitted signal vectors have M_T complex valued components and are selected to send the information signals over the a collection of independent spatial subchannels;

transmitting the sequence of transmitted signal vectors over a spatial channel coupling an array of M_T antenna elements at the base station with an array of M_R antenna elements at the subscriber unit;

receiving a sequence of received signal vectors at the subscriber unit, wherein the received signal vectors have M_R complex valued components; and

performing a space-frequency maximum likelihood detection upon the received signal vectors to recover the information signals.

2. The method of claim 1 wherein the encoding step comprises performing matrix orthogonal frequency division multiplexing of the information signals.

3. The method of claim 1 further comprising the step of injecting training sequences into the information signals.

4. The method of claim 1 further comprising adding cyclic prefixes to the coded signal prior to the transmitting step.

5. The method of claim 1 wherein the encoding step is performed in accordance with a spatio-temporal subchannel decomposition of the channel into independent modes.

6. A digital wireless communication system comprising:

a base station comprising a base station antenna array and a base station signal processor coupled to the base station antenna array;

a subscriber unit comprising a subscriber antenna array coupled through a wireless channel to the base station antenna array and a subscriber signal processor coupled to the subscriber antenna array;

wherein the base station signal processor encodes downlink signal information by matrix orthogonal frequency division multiplexing the signal information; and

wherein the subscriber signal processor decodes the downlink signal information by vector maximum likelihood detection and space-frequency matrix channel estimation.

7. The system of claim 6 wherein the base station signal processor performs interleaving of the signal information and wherein the subscriber signal processor performs de-interleaving.

8. The system of claim 6 wherein the base station signal processor performs trellis encoding of the signal information.

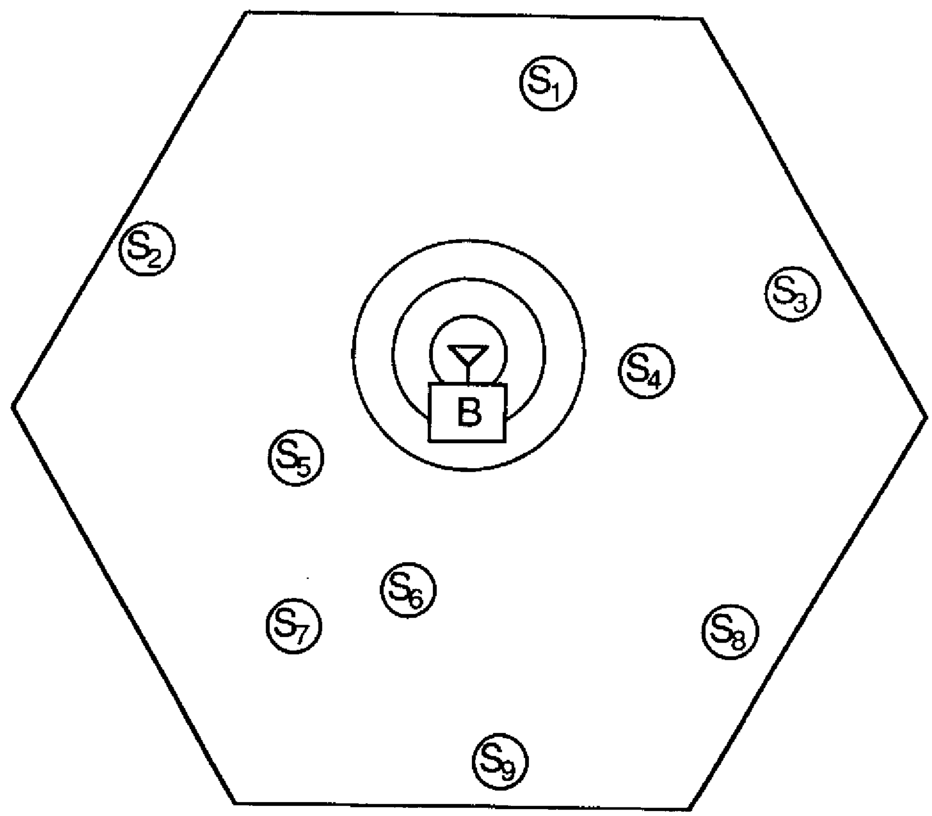

FIG. 1
(PRIOR ART)

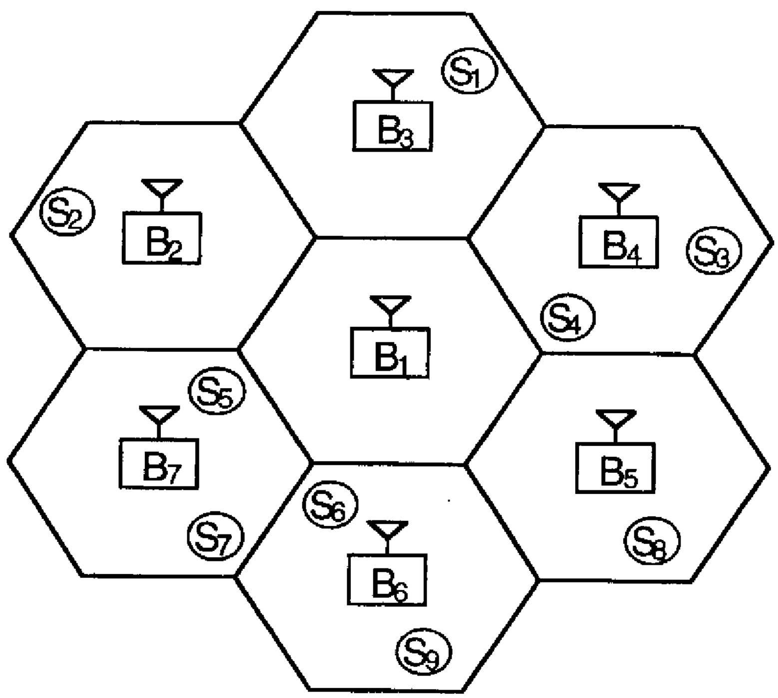

FIG. 2
(PRIOR ART)

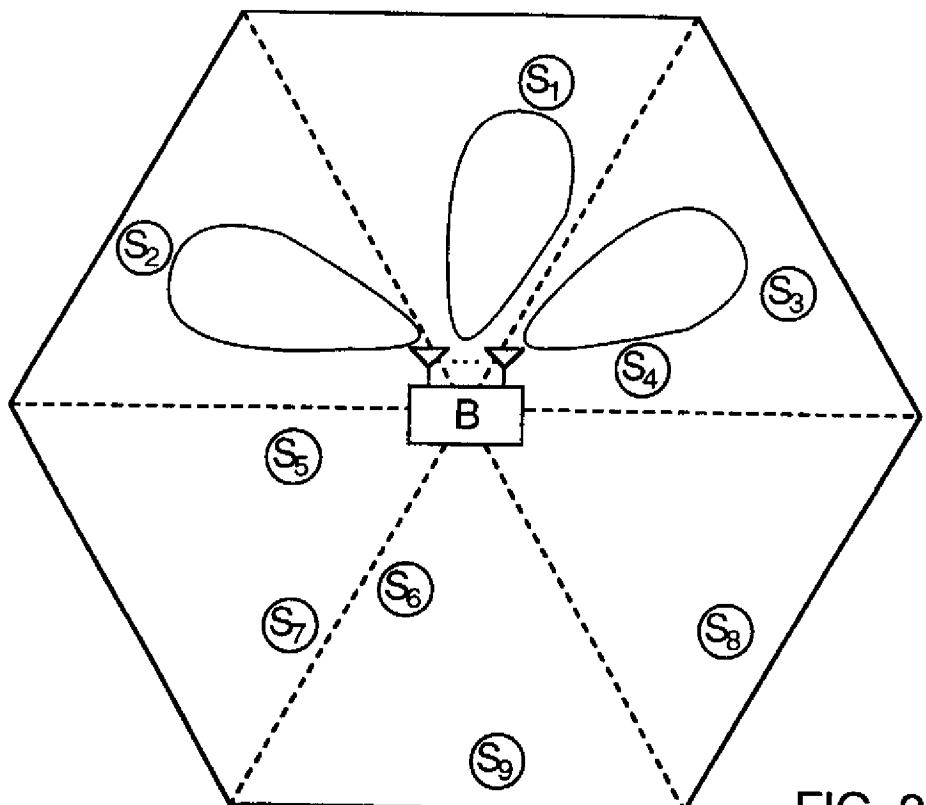

FIG. 3
(PRIOR ART)

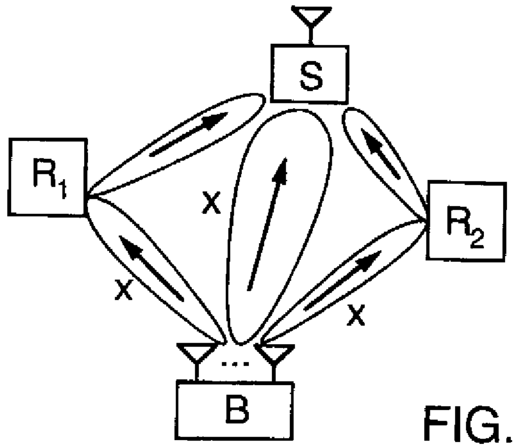

FIG. 4A
(PRIOR ART)

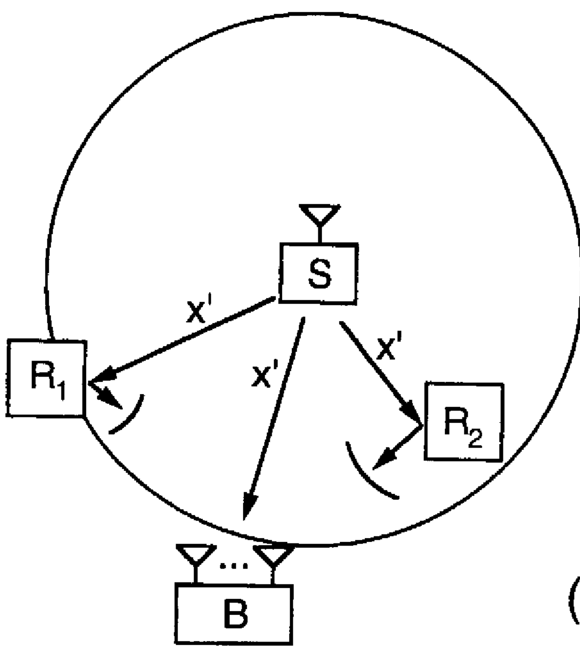

FIG. 4B
(PRIOR ART)

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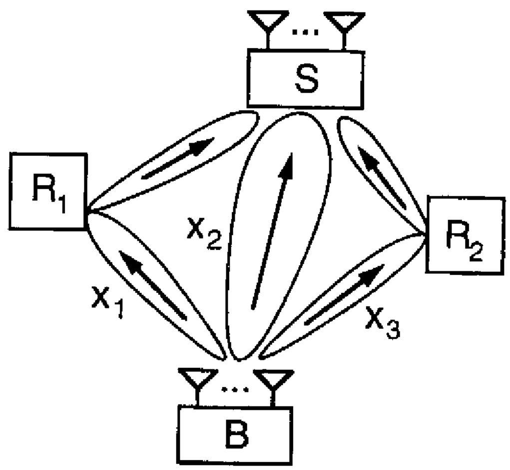


FIG. 5A

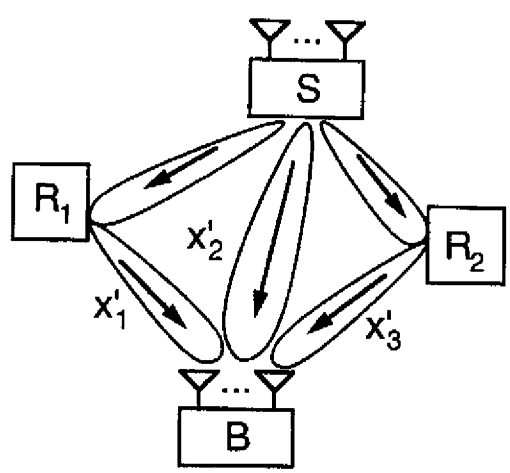
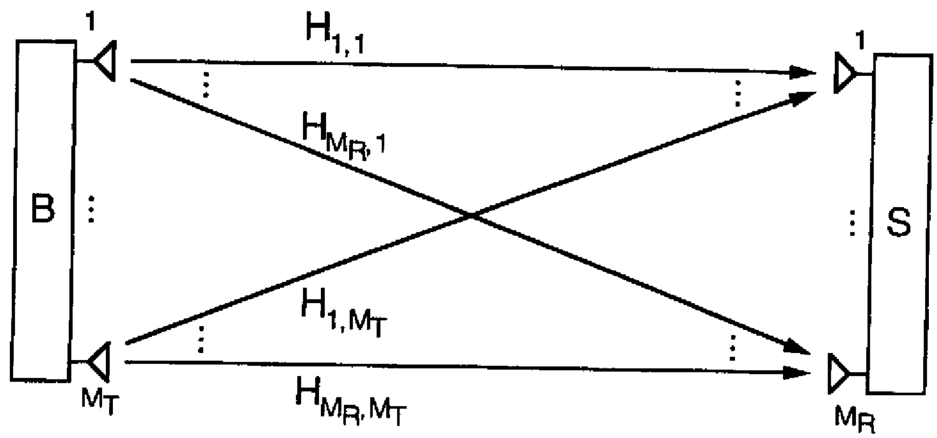
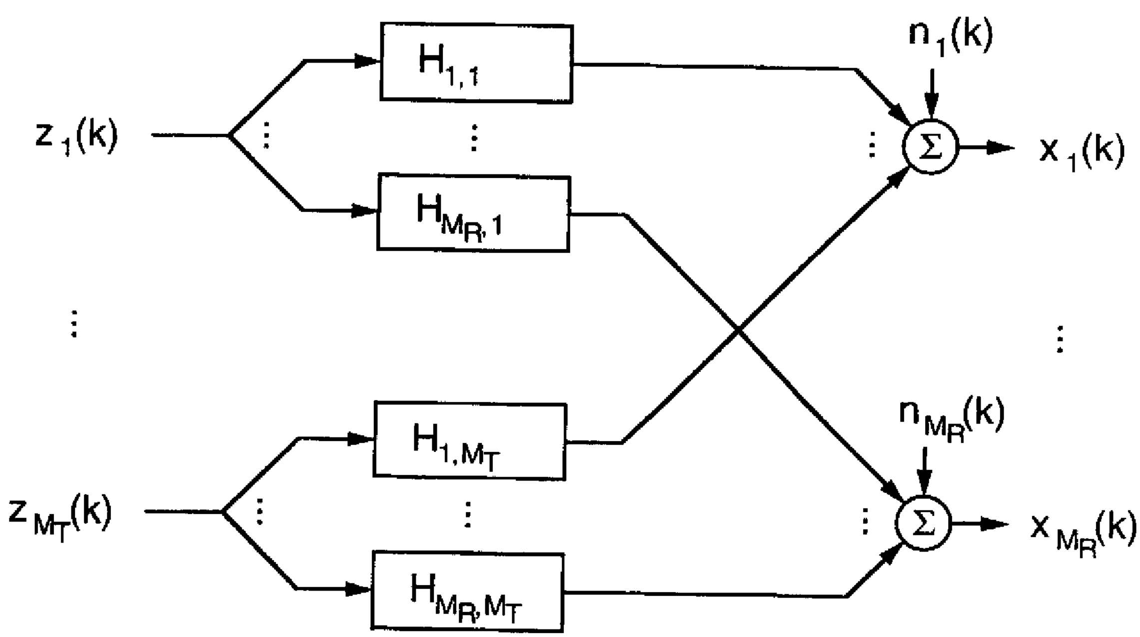


FIG. 5B



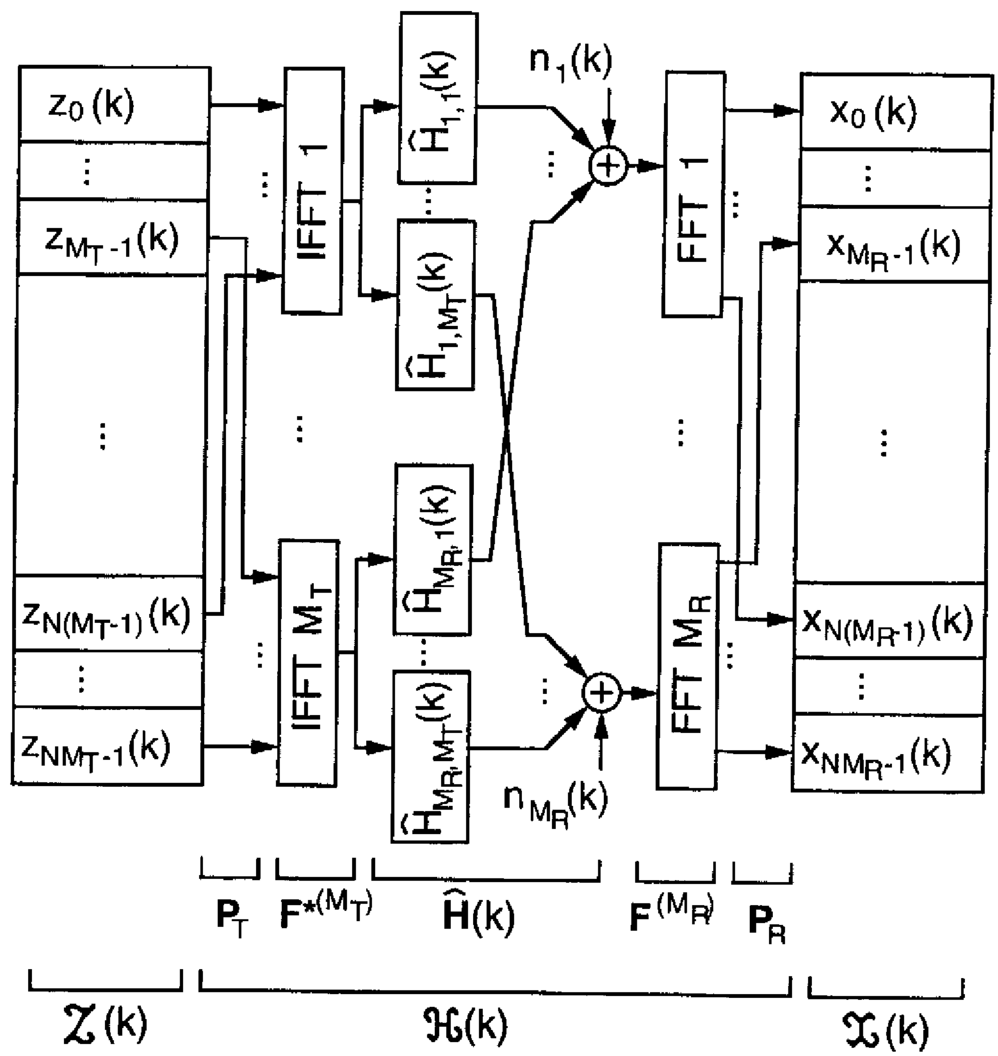
1
H1,1
1
HMR,1
B
S
H1,MT
MT
HMR,MT
MR

FIG. 6A



n1(k)
H1,1
z1(k)
Σ
x1(k)
HMR,1
nMR(k)
H1,MT
zMT(k)
Σ
xMR(k)
HMR,MT

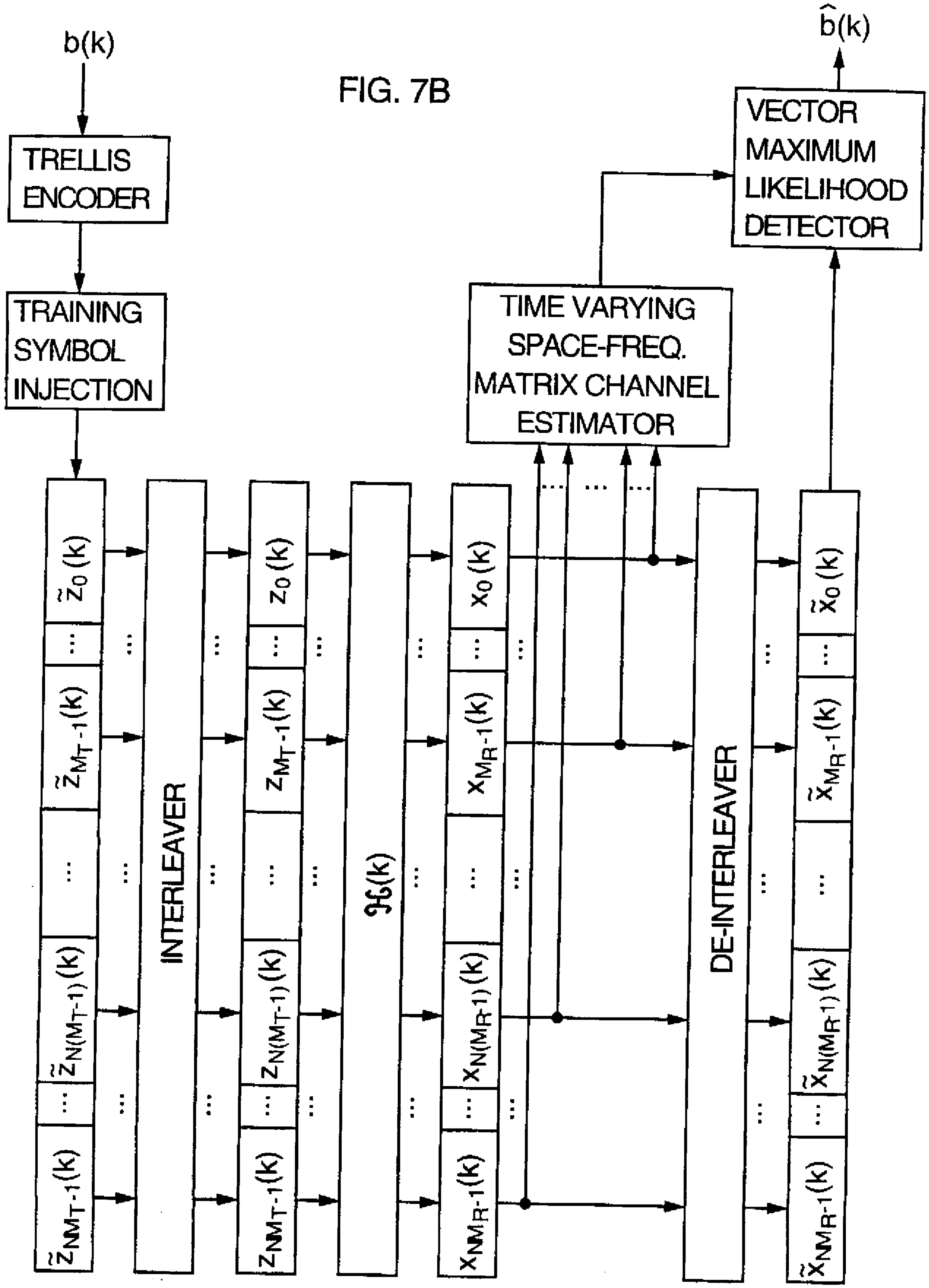
FIG. 6B



$z_0(k)$
$z_{M_T-1}(k)$
$z_{N(M_T-1)}(k)$
$z_{NM_T-1}(k)$
IFFT 1
IFFT M_T
$\hat{H}_{1,1}(k)$
$\hat{H}_{1,M_T}(k)$
$\hat{H}_{M_R,1}(k)$
$\hat{H}_{M_R,M_T}(k)$
$n_1(k)$
$n_{M_R}(k)$
FFT 1
FFT M_R
$x_0(k)$
$x_{M_R-1}(k)$
$x_{N(M_R-1)}(k)$
$x_{NM_R-1}(k)$
$\mathbf{P}_T$
$\mathbf{F}^{*(M_T)}$
$\hat{\mathbf{H}}(k)$
$\mathbf{F}^{(M_R)}$
$\mathbf{P}_R$
$\mathfrak{Z}(k)$
$\mathfrak{H}(k)$
$\mathfrak{X}(k)$

FIG. 7A

FIG. 7B



b(k)
TRELLIS ENCODER
TRAINING SYMBOL INJECTION
$\tilde{z}_0(k)$
$\tilde{z}_{M_T-1}(k)$
$\tilde{z}_{N(M_T-1)}(k)$
$\tilde{z}_{NM_T-1}(k)$
INTERLEAVER
$z_0(k)$
$z_{M_T-1}(k)$
$z_{N(M_T-1)}(k)$
$z_{NM_T-1}(k)$
$\mathcal{H}(k)$
$x_0(k)$
$x_{M_R-1}(k)$
$x_{N(M_R-1)}(k)$
$x_{NM_R-1}(k)$
TIME VARYING SPACE-FREQ. MATRIX CHANNEL ESTIMATOR
DE-INTERLEAVER
$\tilde{x}_0(k)$
$\tilde{x}_{M_R-1}(k)$
$\tilde{x}_{N(M_R-1)}(k)$
$\tilde{x}_{NM_R-1}(k)$
VECTOR MAXIMUM LIKELIHOOD DETECTOR
$\hat{b}(k)$

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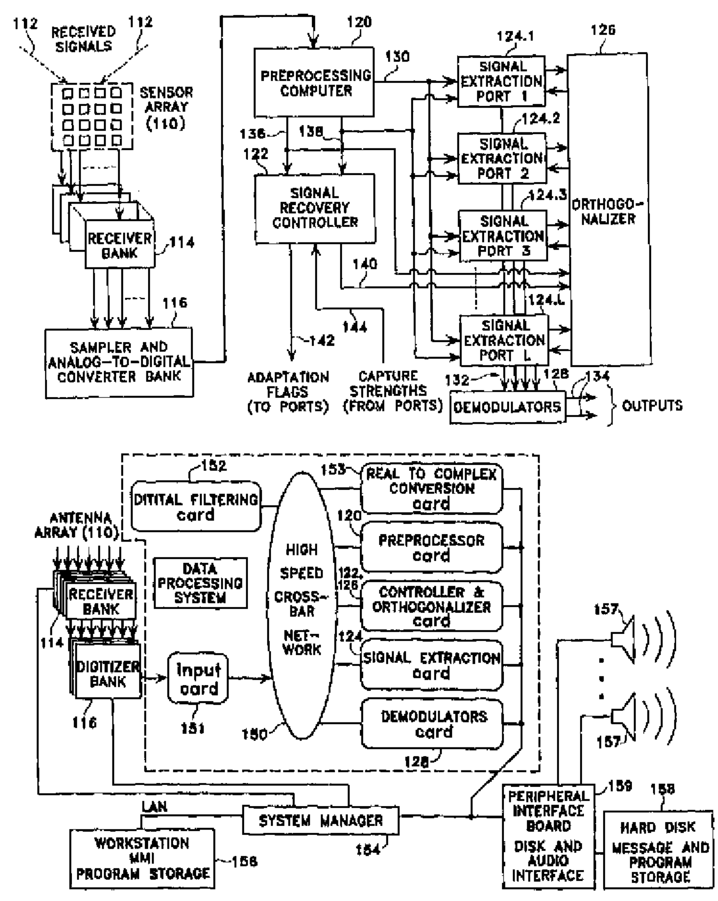
(54) Title: COCHANNEL SIGNAL PROCESSING SYSTEM

(57) Abstract

A method and apparatus for processing cochannel signals received at a sensor array (110) in a cumulant-based signal processing and separation engine to obtain a desired set of output signals (38) or parameters. For use in a signal recovery system, the output signals are recovered and separated versions of the originally transmitted cochannel signals. An important feature that distinguishes the cumulant-based system from other signal separation and recovery systems is that it generates an estimated generalized steering vector associated with each signal source, and representative of all received coherent signal components attributable to the source. This feature enables the invention to perform well in multipath conditions, by combining all coherent multipath components from the same source. In a receiver/transmitter system (316), the estimated generalized steering vectors associated with each source are used to generate transmit beamformer weight vectors that permit cochannel transmission to multiple user stations (310). The basic cumulant-based processing and separation engine can also be used in a variety of applications, such as high density recording, complex phase angle equalization, receiving systems with enhanced effective dynamic range, and signal separation in the presence of strong interference. Various embodiments and extensions of the basic cumulant-based system are disclosed.



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COCHANNEL SIGNAL PROCESSING SYSTEM

BACKGROUND OF THE INVENTION

1. Field of the Invention

This invention relates generally to signal processing systems and, more particularly, to apparatus and methods for receiving and processing signals that share a common receiver frequency band at the same time, referred to as cochannel signals. Even two signals transmitted on slightly separated frequency bands may be "cochannel" signals as seen by a receiver operating to receive signals on a bandwidth that overlaps both of the signals. In a variety of signal processing applications, there is a need to recover information contained in such multiple, simultaneously received signals. In the context of this invention, the word "recover" or "recovery" encompasses separation of the received signals, "copying" the signals (i.e., retrieving any information contained in them), and, in some applications, combining signals received over multiple paths from a single source. The "signals" may be electromagnetic signals transmitted in the atmosphere or in space, acoustic signals transmitted through liquids or solids, or other types of signals characterized by a time-varying parameter, such as the amplitude of a wave. In accordance with another aspect of the invention, signal processing includes transmission of cochannel signals.

In the environment of the present invention, signals are received by "sensors." A sensor is an appropriately selected transducer for converting energy contained in the signal to a more easily manipulated form, such as electrical energy. In a radio communications application, electromagnetic signals are received by antennas and converted to electrical signals for further processing. After separation of the signals, they may be forwarded separately to transducers of a different type, such as loudspeakers, for converting the separated electrical signals into audio signals. In some applications, the signal content may be of less importance than the directions from which the signals were received, and in other applications the received signals may not be amenable to conversion to audible form. Instead, each recovered signal may contain

information in digital form, or may contain information that is best understood by displaying it on a chart or electronic display device. Regardless of the environment in which the present invention is employed, it is characterized by multiple signals received by sensors simultaneously at the same or overlapping frequencies, the need to separate, recover, identify or combine the signals and, optionally, some type of output transducer to put the recovered information in a more easily discernible form.

2. Description of Related Art

Separation and recovery of signals of different frequencies is a routine matter and is handled by appropriate filtering of the received signals. It is common knowledge that television and radio signals are transmitted on different frequency bands and that one may select a desired signal by tuning a receiver to a specific channel. Separation and recovery of multiple signals transmitted at different frequencies and received simultaneously may be effected by similar means, using multiple tuned receivers in parallel. A more difficult problem, and the one with which the present invention is concerned, is how to separate and copy signals from multiple sources when the transmitted signals are at the same or overlapping frequencies. A single sensor, such as an antenna, is unable to distinguish between two or more received signals at the same frequency. However, antenna array technology provides for the separation of signals received from different directions. Basically, and as is well understood by antenna designers, an antenna array can be electronically "steered" to transmit or receive signals to or from a desired direction. Moreover, the characteristics of the antenna array can be selectively modified to present "nulls" in the directions of signals other than that of the signal of interest. A further development in the processing of array signals was the addition of a control system to steer the array toward a signal of interest. This feature is called adaptive array processing and has been known for at least two to three decades. See, for example, a paper by B. Widrow, P.E. Mantey, L.J. Griffiths and B.B. Goode, "Adaptive Antenna Systems," Proceedings of the IEEE, vol. 55, no. 12, pp. 2143-2159, December 1967. The steering characteristics of the antenna can be rapidly switched to receive signals from multiple directions in a

"time-sliced" manner. At one instant the antenna array is receiving a signal from one source and at the next instant, from a different source in a different direction, but information from the multiple sources is sampled rapidly enough to provide a complete record of all the received signals. It will be understood that, although steered antenna array technology was developed principally in the communications and radar fields, it is also applicable to the separation of acoustic and other types of signals.

In the communications field, signals take a variety of forms. Stated most generally, a communication signal typically includes a carrier signal at a selected frequency, on which is impressed or modulated an information signal. There are a large number of different modulation schemes, including amplitude modulation, in which the amplitude of the signal is varied in accordance with the value of an information signal, while the frequency stays constant, and frequency or phase modulation, in which the amplitude of the signal stays constant while its frequency or phase is varied to encode the information signal onto the carrier. Various forms of frequency and phase modulation are often referred to as constant modulus modulation methods, because the amplitude or modulus of the signal remains constant, at least in theory. In practice, the modulus is subject to distortion during transmission, and various devices, such as adaptive equalizers, are used to restore the constant-modulus characteristic of the signal at a receiver. The constant modulus algorithm was developed for this purpose and later applied to antenna arrays in a process called adaptive beam forming The following references are provided by way for further background on the constant modulus algorithm:

B. Agee, "The least-squares CMA: a new technique for rapid correction of constant modulus signals," *Proc. ICASSP-86,* pp. 953-956, Tokyo, Japan, April 1986.

R. Gooch, and J. Lundell, "The CM array, an adaptive beamformer for constant modulus signals," *Proc. ICASSP-86,* pp. 2523-2526, Tokyo, Japan, April 1986.

J. Lundell, and B. Widrow, "Applications of the constant modulus adaptive algorithm to constant and non-constant modulus signals," *Proc. Twenty-*

Second Asilomar Conference on Signals, Systems, and Computers, pp. 432-436, Pacific Grove, CA, November 1988.

B.G. Agee, "Blind separation and capture of communication signals using a multi-target constant modulus beamformer," *Proc. 1989 IEEE Military Communications Conference,* pp. 340-346, Boston, MA, October 1989.

R.D. Hughes, E.H. Lawrence, and L.P. Withers, Jr., "A robust adaptive array for multiple narrowband sources," *Proc. Twenty-Sixth Asilomar Conference on Signals, Systems, and Computers,* pp. 35-39, Pacific Grove, CA, November 1992.

J.J. Shynk and R.P. Gooch, "Convergence properties of the multistage CMA adaptive beamformer," *Proc. Twenty-Seventh Asilomar Conference on Signals, Systems, and Computers,* pp. 622-626, Pacific Grove, CA, November 1993.

The constant modulus algorithm works satisfactorily only for constant modulus signals, such as frequency-modulated (FM) signals or various forms of phase-shift keying (PSK) in which the phase is discretely or continuously varied to represent an information signal, but not for amplitude-modulated (AM) signals or modulation schemes that employ a combination of amplitude and phase modulation. There is a significant class of modulation schemes used known as M-ary quadrature amplitude modulation (QAM), used for transmitting digital data, whereby the instantaneous phase and amplitude of the carrier signal represents a selected data state. For example, 16-ary QAM has sixteen distinct phase-amplitude combinations. The "signal constellation" diagram for such a scheme has sixteen points arranged in a square matrix and lying on three separate constant-modulus circles. A signal constellation diagram is a convenient way of depicting all the possible signal states of a digitally modulated signal. In such a diagram, phase is represented by angular position and modulus is represented by distance from an origin.

The constant modulus algorithm has been applied with limited success to a 16-ary QAM scheme, because it can be represented as three separate constant-modulus signal constellations. However, for higher orders of QAM the constant modulus algorithm provides rapidly decreasing accuracy. For suppressed-carrier AM,

the constant modulus approach fails completely in trying to recover cochannel AM signals. If there are multiple signals, the constant modulus algorithm yields signals with "cross-talk," i.e. with information in the two signals being confused. For a single AM signal in the presence of noise, the constant-modulus algorithm yields a relatively noisy signal.

Because antenna arrays can be steered electronically to determine the directions of signal sources, it was perhaps not surprising that one well known form signal separator available prior to the present invention used direction finding as its basis. The approach is referred to as DF-aided copy, where DF means direction finding. This is an open-loop technique in which steering vectors that correspond to estimated signal source bearings are first determined; then used to extract waveforms of received signals. However, the direction finding phase of this approach requires a knowledge of the geometry and performance characteristics of the antenna array. Then steering vectors are fed forward to a beamformer, which nulls out the unwanted signals and steers one or more antenna beam(s) toward each selected source.

Prior to the present invention, some systems for cochannel signal separation used direction-finding (DF)-beamforming. Such systems separate cochannel signals by means of a multi-source (or cochannel) super-resolution direction finding algorithm that determines steering vectors and directions of arrival (DOAs) of multiple simultaneously detected cochannel signal sources. An algorithm determines beam-forming weight vectors from the set of steering vectors of the detected signals. The beamforming weight vectors are then used to recover the signals. Any of several well-known multi-source super-resolution DF algorithms can be used in such a system. Some of the better known ones are usually referred to by the acronyms MUSIC (MUltiple SIgnal Classification), ESPRIT (Estimation of Signal Parameters via Rotational Invariance Techniques), Weighted Subspace Fitting (WSF), and Method of Direction Estimation (MODE).

MUSIC was developed in 1979 simultaneously by Ralph Schmidt in the United States and by Georges Bienvenu and Lawrence Kopp in France. The Schmidt work is described in R.O. Schmidt, "Multiple emitter location and signal parameter

estimation," *Proc. RADC Spectrum Estimation Workshop,* pp. 243-258, Rome Air Development Center, Griffiss Air Force Base, NY, October 3-5, 1979. The Bienvenu work is described in G. Bienvenu and L. Kopp, "Principe de la goniometrie passive adaptative," *Proc. Colloque GRETSI,* pp. 106/1-106/10, Nice, France, May 1979. MUSIC has been extensively studied and is the standard against which other super-resolution DF algorithms are compared.

ESPRIT is described in many publications in the engineering signal processing literature and is the subject of United States Patent No. 4,750,147 entitled "Method for estimating signal source locations and signal parameters using an array of sensor pairs," issued to R.H. Roy III et al. ESPRIT was developed by Richard Roy, III, Arogyaswami Paulraj, and Prof. Thomas Kailath at Stanford University. It was presented as a super-resolution algorithm for direction finding in the following series of publications starting in 1986:

A. Paulraj, R. Roy, and T. Kailath, "A subspace rotation approach to signal parameter estimation," Proc. IEEE, vol. 74, no. 4, pp. 1044-1045, July 1986.

R. Roy, A. Paulraj, and T. Kailath, "ESPRIT - A subspace rotation approach to estimation of parameters of cisoids in noise," IEEE Trans. Acoust., Speech, and Signal Processing, vol. ASSP-34, no. 5, pp. 1340-1342, October 1986.

R.H. Roy, ESPRIT - Estimation of Signal Parameters via Rotational Invariance Techniques, doctoral dissertation, Stanford University, Stanford, CA, 1987.

R. Roy and T. Kailath, "ESPRIT - Estimation of signal parameters via rotational invariance techniques," IEEE Trans. Acoust., Speech, and Signal Processing, vol. ASSP-37, no. 7, pp. 984-995, July 1989.

B. Ottersten, R. Roy, and T. Kailath, "Signal waveform estimation in sensor array processing," Proc. Twenty-Third Asilomar Conference on Signals, Systems, and Computers, pp. 787-791, Pacific Grove, CA, November 1989.

R. Roy and T. Kailath, "ESPRIT - Estimation of signal parameters via rotational invariance techniques," Optical Engineering, vol. 29, no. 4, pp. 296-313, April 1990.

MUSIC and ESPRIT both require the same "narrowband array assumption," which is further discussed below in the detailed description of the invention, and both are modulation independent, a feature shared by all cochannel signal separation and recovery techniques that are based on the DF-beamforming method.

ESPRIT calculates two N-by-N covariance matrices, where N is the number of antenna elements, and solves a generalized eigenvalue problem numerically (instead of using a calibration table search, as MUSIC does). It does this for every block of input samples. MUSIC calculates a single N-by-N covariance matrix, performs an eigendecomposition, and searches a calibration table on every block of input array samples (snapshots).

MUSIC and ESPRIT have a number of shortcomings, some of which are discussed in the following paragraphs.

ESPRIT was successfully marketed based on a single, key advantage over MUSIC. Unlike MUSIC, ESPRIT did not require array calibration. In ESPRIT, the array calibration requirement was eliminated, and a different requirement on the antenna array was substituted. The new requirement was that the array must have a certain geometrical property. Specifically the array must consist of two identical sub-arrays, one of which is offset from the other by a known displacement vector. In addition, ESPRIT makes the assumption that the phases of received signals at one sub-array are related to the phases at the other sub-array in an ideal theoretical way.

Another significant disadvantage of ESPRIT is that, although it purports not to use array calibration, it has an array manifold assumption hidden in the theoretical phase relation between sub-arrays. "Array manifold" is a term used in antenna design to refer to a multiplicity of physical antenna parameters that, broadly speaking, define the performance characteristics of the array.

A well known difficulty with communication systems, especially in an urban environment, is that signals from a single source may be received over multiple paths that include reflections from buildings and other objects. The multiple paths may interpose different time delays, phase changes and amplitude changes on the

transmitted signals, rendering reception more difficult, and transmission uncertain. This difficulty is referred to as the multipath problem. It is one that has not been adequately addressed by signal processing systems of the prior art.

Neither MUSIC nor ESPRIT can operate in a coherent multipath environment without major added complexity. A related problem is that, in a signal environment devoid of coherent multipath, no DF-beamforming method can separate signals from sources that are collinear with the receiving array, i.e. signal sources that are in line with the array and have zero angular separation. Even in a coherent multipath environment, DF-beamforming methods like MUSIC and ESPRIT cannot separate and recover cochannel signals from collinear sources.

Another difficulty with ESPRIT is that it requires two antenna sub-arrays and is highly sensitive to mechanical positioning of the two sub-arrays, and to the electromagnetic matching of each antenna in one sub-array with its counterpart in the other sub-array. Also ESPRIT requires a 2N-channel receiver, where N is the number of antenna elements, and is highly sensitive to channel matching.

Another significant drawback in both MUSIC and ESPRIT is that they fail abruptly when the number of signals detected exceeds the capacity, N, equal to the number of antennas in the case of MUSIC, or half the number of antennas in the case of ESPRIT.

A fundamental problem with both MUSIC and ESPRIT is that they use open-loop feed-forward computations, in which errors in the determined steering vectors are uncorrected, uncorrectable, and propagate into subsequent calculations. As a consequence of the resultant inaccurate steering vectors, MUSIC and ESPRIT have poorcross-talk rejection, as measured by signal-to-interference-plus-noise ratio (SINR) at the signal recovery output ports.

ESPRIT is best suited to ground based systems where its antenna requirements are best met and significant computational resources are available. MUSIC has simpler antenna array requirements and lends itself to a wider range of platforms, but also needs significant computational resources.

Another limitation of most signal recovery systems of the prior art is that they rely on first-order and second-order statistical moments of the received signal data. A moment is simply a statistical quantity derived from the original data by mathematical processing at some level. An average or mean value of the several signals received at a given time is an example of a first-order moment. The average of the squares of the signal values (proportional to signal powers) is an example of a second-order moment. Even if one considers just one signal and a noise component, computing the average of the sum of the squares produces a cross-term involving the product of signal and noise components. Typically, engineers have managed to find a way to ignore the cross-term by assuming that the signal and the noise components are statistically independent. At a third-order level of statistics, one has to assume that the signal and noise components have zero mean values in order to eliminate the cross-terms in the third-order moment. For the fourth-order and above, the computations become very complex and are not easily simplified by assumptions. In most prior art signal analysis systems, engineers have made the gross assumption that the nature of all signals is Gaussian and that there is no useful information in the higher-order moments. Higher-order statistics have been long recognized in other fields and there is recent literature suggesting their usefulness in signal recovery. Prior to this invention, cumulant-based solutions have been proposed to address the "blind" signal separation problem, i.e. the challenge to recover cochannel signals without knowledge of antenna array geometry or calibration data. See, for example, the following references:

J.-F. Cardoso, "Source separation using higher order moments," *Proc. ICASSP-89*, pp. 2109-2112, Glasgow, Scotland, May 1989.

J.-F. Cardoso, "Eigen-structure of the fourth-order cumulant tensor with application to the blind source separation problem," *Proc. ICASSP-90*, pp. 2655-2658, Albuquerque, New Mexico, April 1990.

J.-F. Cardoso, "Super-symmetric decomposition of the fourth-order cumulant tensor: blind identification of more sources than sensors," *Proc. ICASSP-91*, pp. 3109-3112, Toronto, Canada, May 1991.

J.-F. Cardoso, "Higher-order narrowband array processing," *International Signal Processing Workshop on Higher Order Statistics*, pp. 121--130, Chamrousse-France, July 10-12, 1991.

J.-F. Cardoso, "Blind beamforming for non-Gaussian sources," *IEE Proceedings Part F*, vol. 140, no. 6, pp. 362--370, December 1993.

P. Comon, "Separation of stochastic processes," *Proc. Vail Workshop on Higher-Order Spectral Analysis*, pp. 174--179, Vail, Colorado, USA, June 1989.

P. Comon, "Independent component analysis," *Proc. of Intl. Workshop on Higher-Order Statistics*, pp. 111-120, Chamrousse, France, 1991.

P. Comon, C. Jutten, and J. Herault, "Blind separation of sources, part II: problems statement," *Signal Processing*, vol. 24, no. 1, pp. 11-20, July 1991.

E. Chaumette, P. Comon, and D. Muller, "ICA-based technique for radiating sources estimation: application to airport surveillance," *IEE Proceedings Part F*, vol. 140, no. 6, pp. 395--401, December 1993.

Z. Ding, "A new algorithm for automatic beamforming," *Proc. Twenty-Fifth Asilomar Conference on Signals, Systems, and Computers*, pp. 689-693, Pacific Grove, CA, November 1991.

M. Gaeta and J.-L. Lacoume, "Source separation without a-priori knowledge: the maximum likelihood solution," *Proc. EUSIPCO*, pp. 621-624, 1990.

E. Moreau, and O. Macchi, "New self-adaptive algorithms for source separation based on contrast functions," *Proc. IEEE SP Workshop on Higher-Order Statistics*, pp. 215-219, Lake Tahoe, USA, June 1993.

P. Ruiz, and J.L. Lacoume, "Extraction of independent sources from correlated inputs: a solution based on cumulants," *Proc. Vail Workshop on Higher-Order Spectral Analysis*, pp. 146--151, Vail, Colorado, USA, June 1989.

E.H. Satorius, J.J. Mulligan, Norman E. Lay, "New criteria for blind adaptive arrays," *Proc. Twenty-Seventh Asilomar Conference on Signals, Systems, and Computers*, pp. 633-637, Pacific Grove, CA, November 1993.

L. Tong, R. Liu, V. Soon, and Y. Huang, "Indeterminacy and identifiability of blind identification," *IEEE Trans. Circuits and Systems*, vol. 38, pp. 499-509, May 1991.

L. Tong, Y. Inouye and R. Liu, "Waveform preserving blind estimation of multiple independent sources," *IEEE Trans. Signal Processing*, vol. 41, no. 7, pp. 2461--2470, July 1993.

However, all of these approaches to blind signal recovery address the static case in which a batch of data is given to a processor, which then determines the steering vectors and exact waveforms. These prior approaches do not have the ability to identify new sources that appear or existing sources that are turned off. In addition, previously proposed algorithms require multiple levels of eigendecomposition of array covariance and cumulant matrices. Their convergence to reliable solutions depends on the initialization and utilization of the cumulant matrices that can be derived from array measurements. Furthermore, previous cumulant-based algorithms have convergence problems in the case of identically modulated sources in general.

Ideally, a system for receiving and processing multiple cochannel signals should make use of statistics of the measurements, and should not need to rely on knowledge of the geometry or array manifold of the sensors, i.e., the array calibration data. Also, the system should be able to receive and process cochannel signals regardless of their modulation or signal type, e.g. it should not be limited to constant-modulus signals. More generally, the ideal cochannel signal processing system should not be limited to any modulation properties, such as baud rate or exact center frequency. Any system that is limited by these properties has only a limited range of source types that can be separated, and is more suitable for interference suppression in situations where the desired signal properties are well known. Another desirable property of the ideal cochannel signal receiving and processing system is that it should operate in a dynamic way, identifying new signal sources that appear and identifying sources that disappear. Another desirable characteristic is a very high speed of operation allowing received signals to be processed in real time. As will shortly

become apparent, the present invention meets and exceeds these ideal characteristics for cochannel signal processing.

SUMMARY OF THE INVENTION

The present invention resides in a system or method for processing cochannel signals received at a sensor array and producing desired recovered signals or parameters as outputs. In the context of this specification, "cochannel" signals are that overlap in frequency, as viewed from a receiver of the signals. Even signals that are transmitted in separate, but closely spaced, frequency bands may be cochannel signals as viewed from a receiver operating in bandwidth wide enough to overlap both of the signals. A key aspect of the invention is that it is capable of separating and recovering multiple cochannel signals very rapidly using only sensor array signals, without knowledge of sensor array geometry and array manifold, (e.g. array calibration data), and without regard to the signal type or modulation. If array calibration data are available, the system also provides direction-of-arrival parameters for each signal source. The invention inherently combines coherent multipath components of a received signal and as a result achieves improved performance in the presence of multipath. One embodiment of the invention also includes a transmitter, which makes use of estimated generalized steering vectors generated while separating and recovering received signals, in order to generate appropriate steering vectors for transmitted signals, to ensure that transmitted signals intended for a particular signal source traverse generally the same path or paths that were followed by signals received from the same signal source.

Briefly, and in general terms, the system of the invention comprises a signal receiving system, including means for generating a set of conditioned receiver signals from cochannel signals of any modulation or type received at a sensor array from multiple sources that can vary in power and location; an estimated generalized steering vector (EGSV) generator, for computing for each source an EGSV that results in optimization of a utility function that depends on fourth or higher even-order